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The Proceedings

OF

THE INSTITUTION OF ELECTRICAL ENGINEERS

FOUNDED 1871 : INCORPORATED BY ROYAL CHARTER 1921

PART B

RADIO AND ELECTRONIC ENGINEERING
(INCLUDING COMMUNICATION ENGINEERING)

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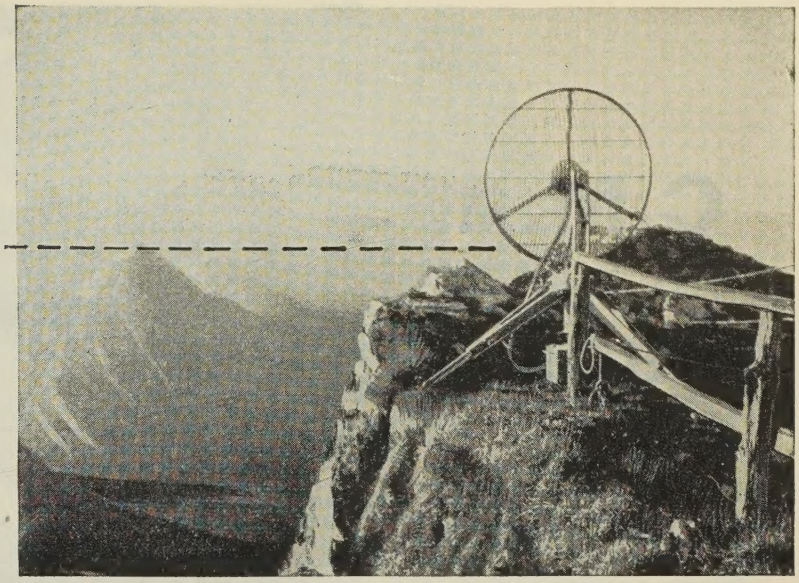
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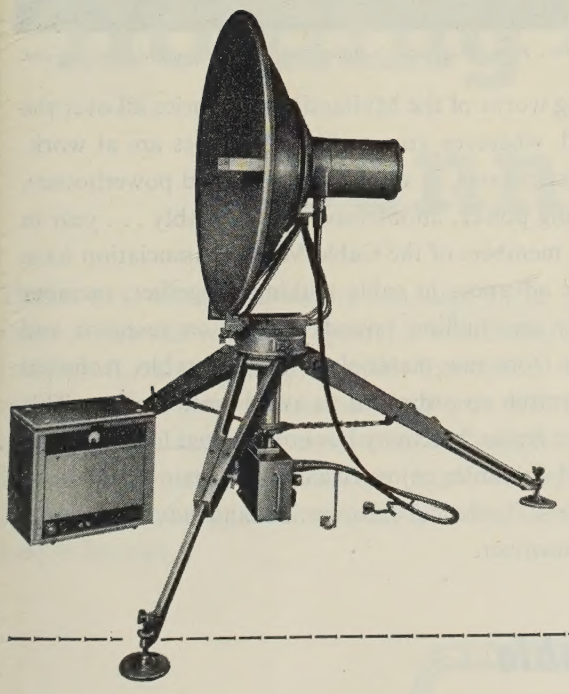
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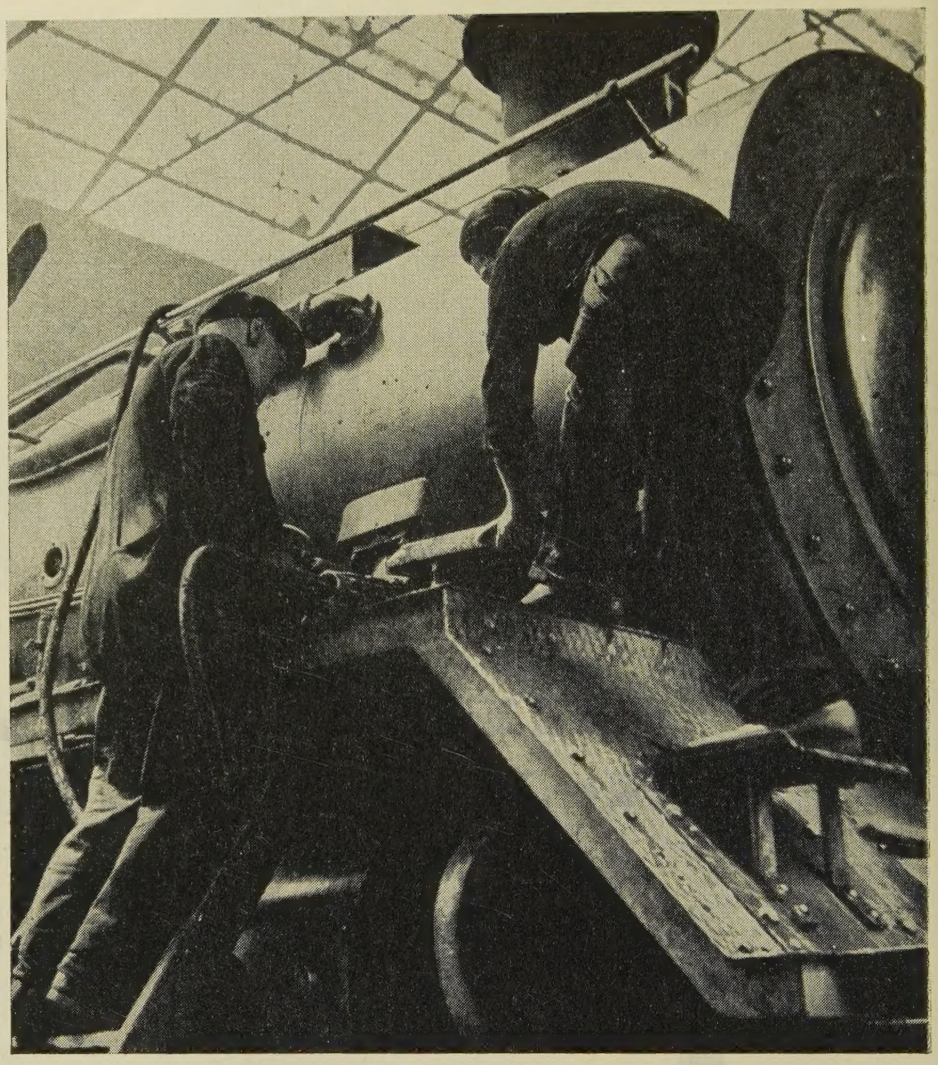
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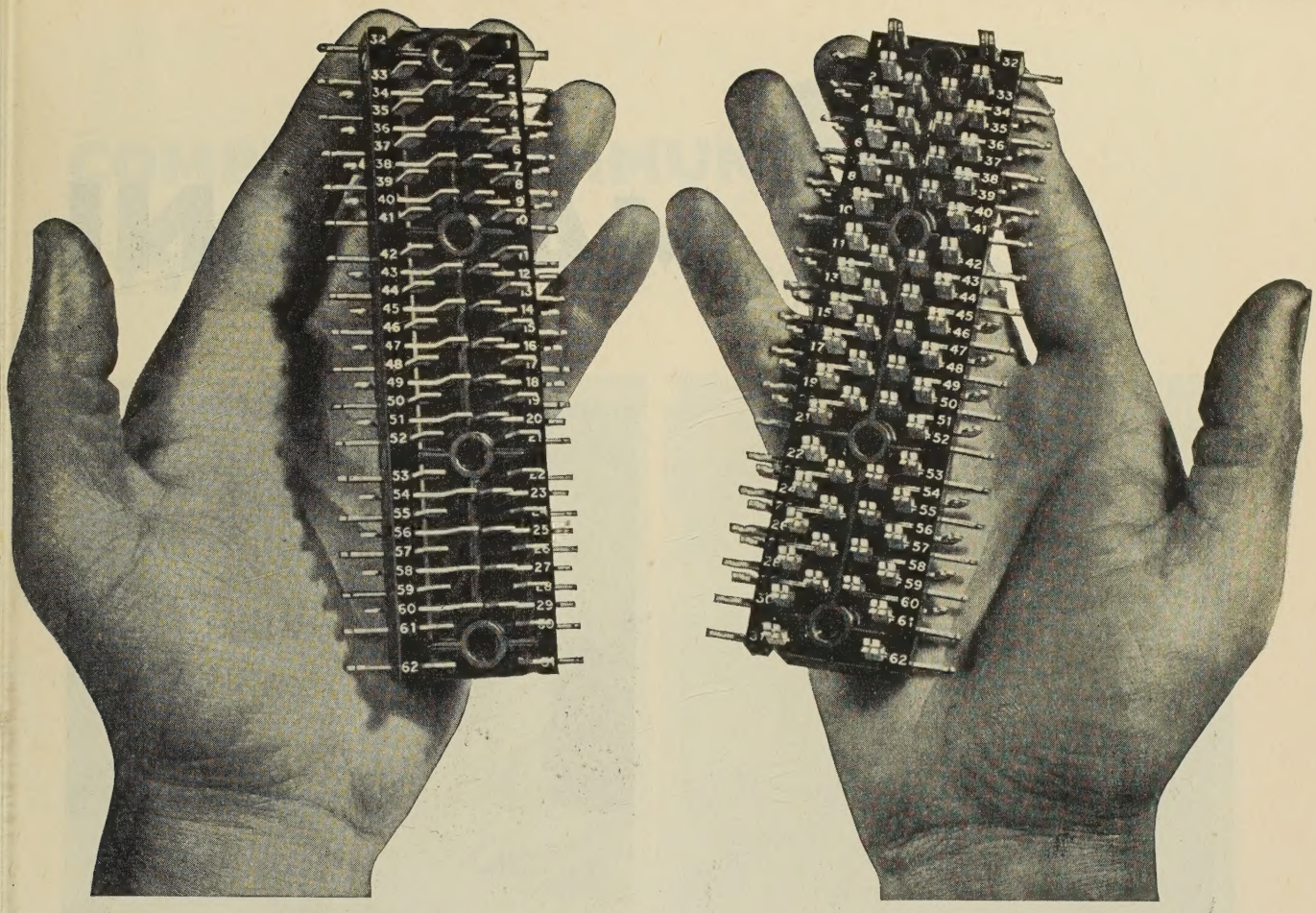
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There are almost certainly enough high grade ones in this 62-way connector. It was designed to fulfil an Admiralty requirement that could not be met by any other connector, and you will find it suitable for many rack-mounting applications where low contact resistance, high voltage and current capacity and excellent insulation are major factors.

SPECIFICATION

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Insulation:	2kV. between contacts
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Current rating:	10A. max. per single contact
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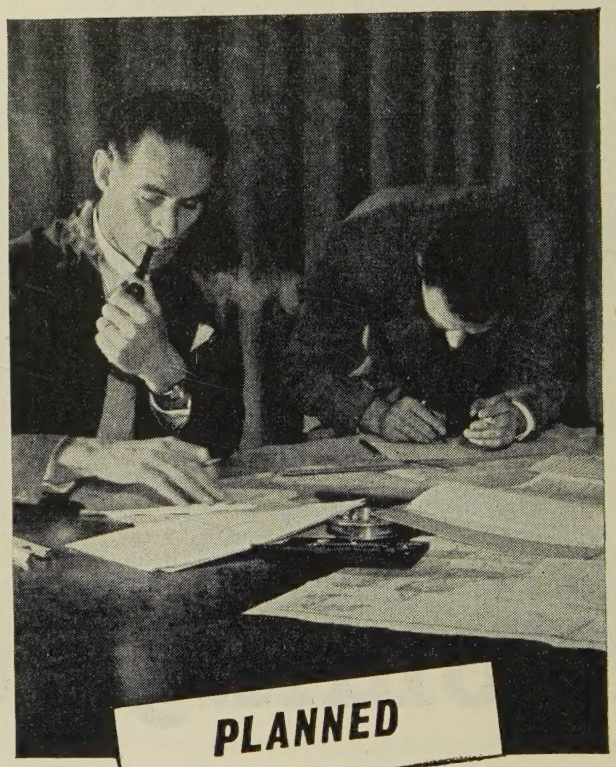
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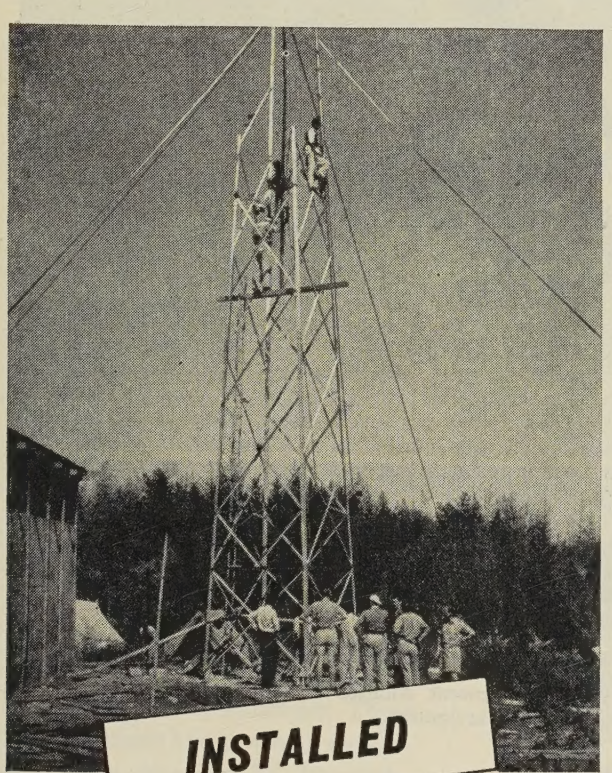
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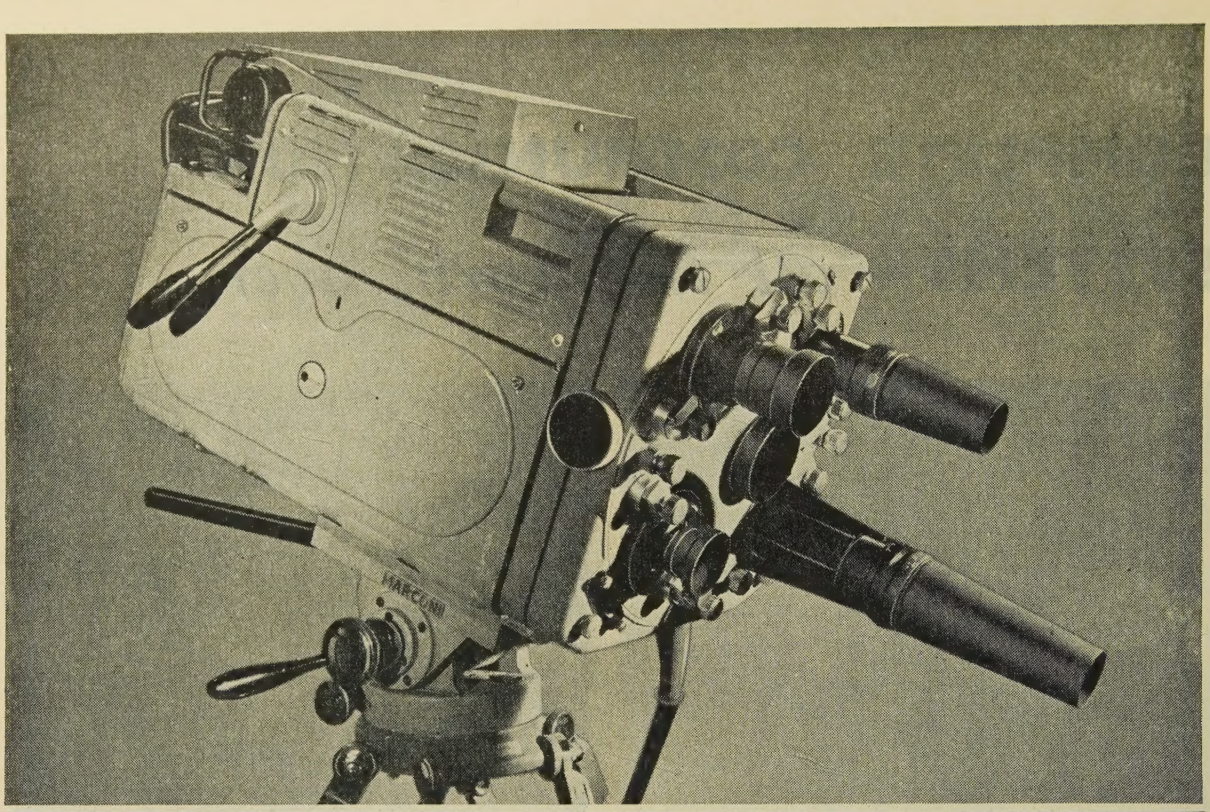
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- Designed for ease of servicing, excellent accessibility and plug-in sub-units.
- Four position turret will carry any combination from 2-inch to 40-inch lenses. 80-inch and zoom lenses may also be used.
- Viewfinder can be tilted up or down to give the most comfortable viewing position.
- Camera Control Unit may be used with 10″ picture tube and 3″ waveform tube, or with 14″ picture tube and 5″ waveform tube.
- Remote control of light intensity by variable graded filter.
- Optional remote control of focus and turret. Optional semi-automatic alignment circuit.
- Built-in turret for neutral density and colour filters.
- Full range of accessories available for both studio and outside broadcast roles.

BROADCAST VIDICON CAMERA Type BD864

The most recent addition to the Marconi range of Television Equipment.

Features

- Compact, easily operated by one man. The camera has integral viewfinder with 7″ tube and 2¾″ waveform monitor and includes all operational controls.
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- Use of close-tolerance double-triodes in all valve circuits except one and printed wiring assemblies ensures great reliability.
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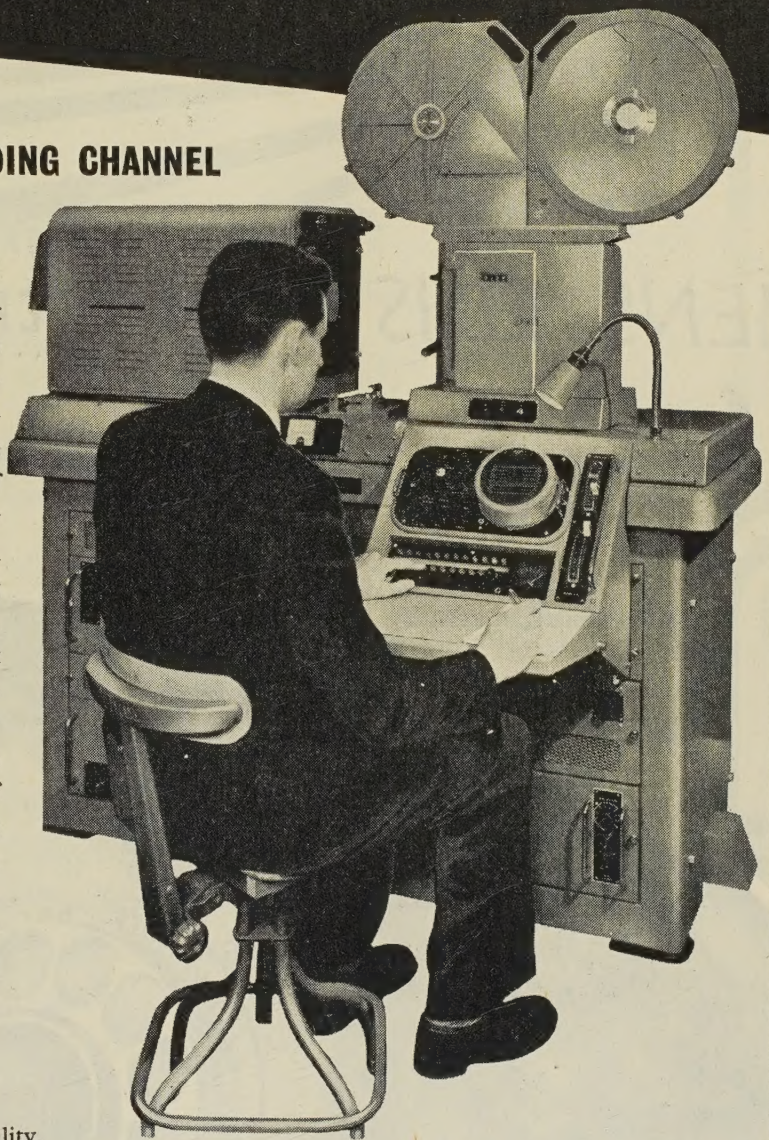
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MARCONI TELEVISION RECORDING EQUIPMENT

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Features

Exceptionally high picture quality.

Specially developed gearbox enables Fast Pull-down technique to be employed. F.P.D. Mechanism has given over 3,000 hours trouble-free operation.

Pull-down time adjustable, normally set at 2 milliseconds permitting recording of fully interlaced picture. Simple single-lens optical system avoids loss of contrast.

Sound can be recorded on optical track, magnetic stripe or separate synchronous magnetic track.

Conveniently placed input selector switches, monitor and level controls. Sound/Vision cueing device incorporated.

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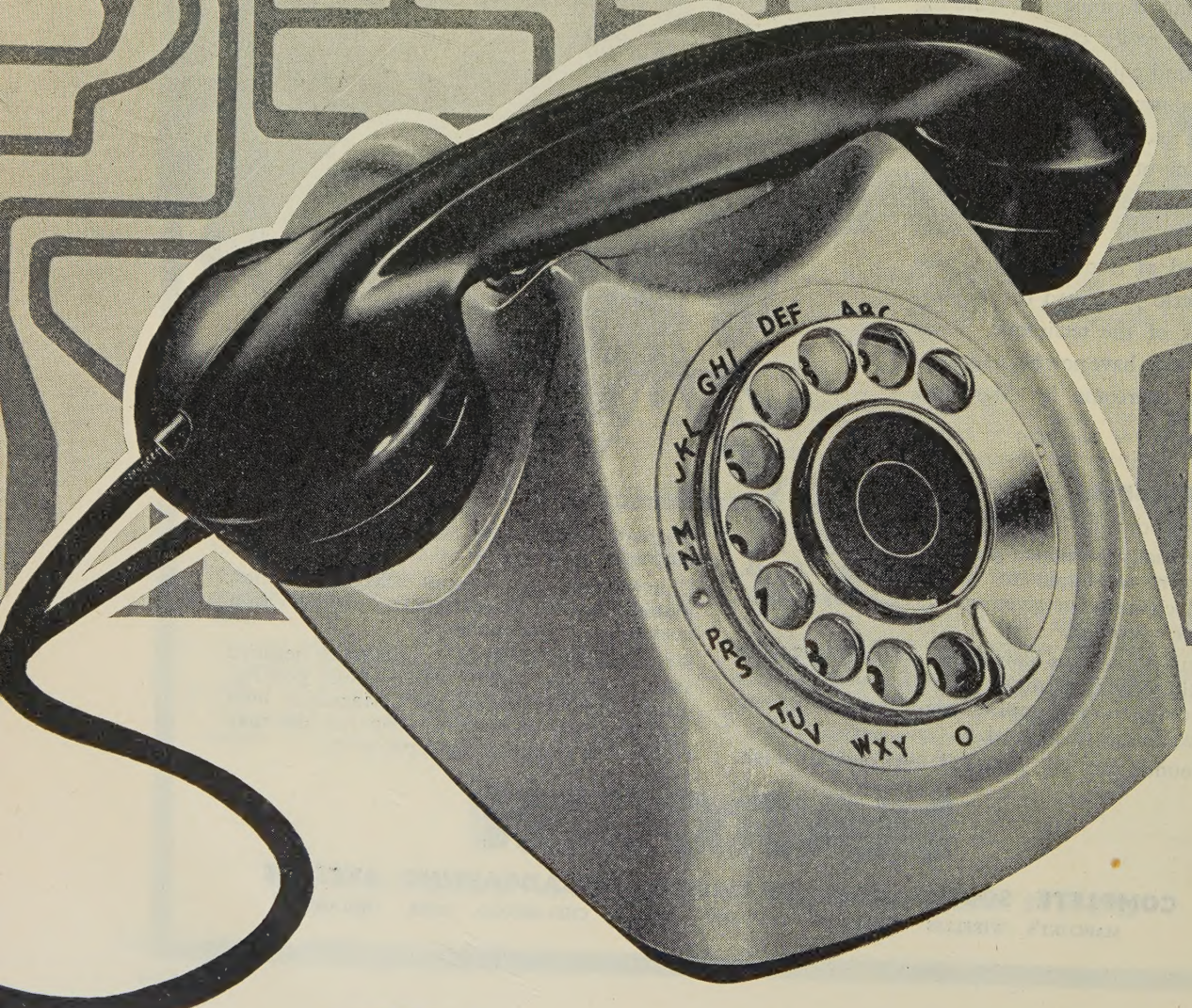
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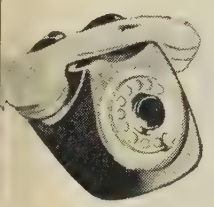
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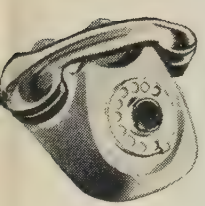
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This entirely new instrument, a product of the designers of the world-famous "Neophone" which established new standards in telephone performance and appearance in the nineteen-thirties, incorporates the following attractive features —



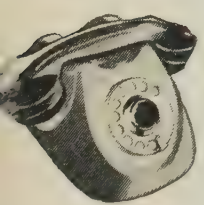
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Reduced size and weight – handset weight halved – only 7 ounces.



21 COLOUR COMBINATIONS

7 different handset shades and a choice of 3 case colours. The case can be changed WITHOUT DISTURBING THE DIAL SWITCH.

The centenary NEOPHONE



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A multi-channel telephone link extensible to 720 speech circuits

G.E.C.

240-circuit U.H.F. RADIO EQUIPMENT

Designed for long life and easy maintenance

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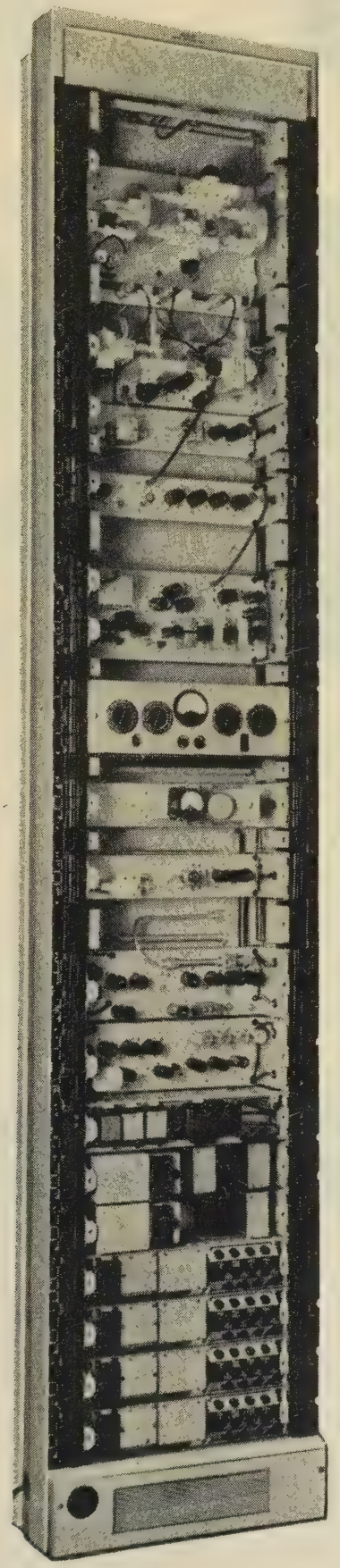
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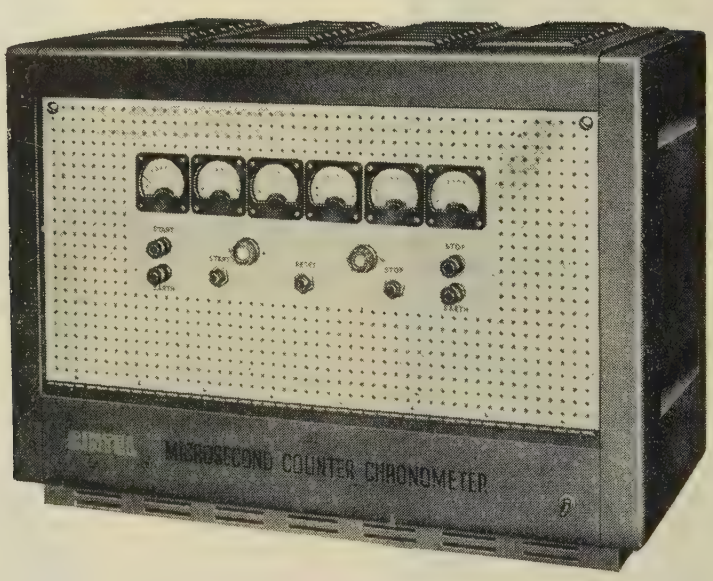
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This equipment is designed to measure small intervals of time to a high order of accuracy and two ranges are provided:

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These have been in regular quantity production for the past two years, and have proved themselves reliable and stable in a *variety* of applications. They are admirably suitable for all forms of DC to DC or DC to AC Converters, High Power portable Amplifiers and Public Address Equipment. "GOLTOP" Power Transistors are the first to be offered for immediate delivery in quantity. Representing the latest developments in semi-conductor technique for power applications, these entirely British-made p-n-p Germanium Junction Transistors will open up entirely new fields to designers of industrial, commercial and military equipment.

Available in 6 TYPES, all for 10-watts power dissipation:
V15/10P. V15/20P. V15/30P. for 15 volts max.
V30/10P. V30/20P. V30/30P. for 30 volts max.

Maximum Collector Power Dissipation (DC or Mean) for all types	$t_{amb}=25°C$	$t_{amb}>25°C$ *Reduction/°C*
(1) Clamped directly on to 50 sq. in. of 16 S.W.G. aluminium	*10W*	*200mW*
(2) Clamped directly on to 9 sq. in. of 16 S.W.G. aluminium	*4W*	*80mW*
(3) As (2) but with 2 mil mica washer between heat sink and transistor	*2W*	*40mW*
(4) Transistor only in free air	*1W*	*20mW*

- **High power rating—up to 10W at audio and supersonic frequencies.**
- **High current ratings up to 3A DC.**
- **Long life.**
- **Excellent resistance to mechanical shock.**
- **Hermetic sealing and rigorous manufacturing control ensure uniformity and stability of a high order.**



British Design, Materials and Craftsmanship

Data sheets gladly forwarded on request

All trade enquiries to: **Newmarket Transistor Co. Ltd.**

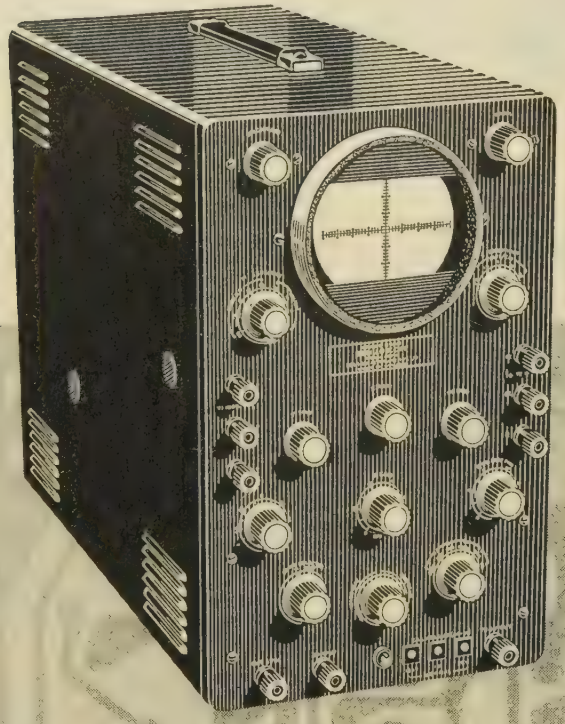
Exning Road, Newmarket. Telephone: Newmarket 3381/4

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Model 1071K ***Double Beam*** *Kit Oscilloscope*
List Price £57.10.0

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Q. *Why has Cossor Instruments decided upon this innovation?*

A. To make available a range of first-class measuring instruments **at a considerable saving in cost to the Buyer.**

Q. *Are Kit instruments inferior in performance to their Factory-built equivalents?*

A. Certainly not. If assembled and wired exactly in accordance with the Manual of Instructions.

Q. *A certain skill must, surely, be required to build these instruments?*

A. None beyond the ability to use a small soldering iron.

Q. *How can a performance specification be maintained without setting up with test equipment?*

A. Largely by the use of PRINTED CIRCUITS which allow no interference with the layout of critical parts of the circuit.

Q. *How many Kit instruments are at present available?*

A. Three. Two Oscilloscopes, a Single-Beam and a Double-Beam, and a Valve Voltmeter. Others will follow shortly.

Q. *Could I have more information on these interesting instruments?*

A. With the greatest of pleasure. Just write to:

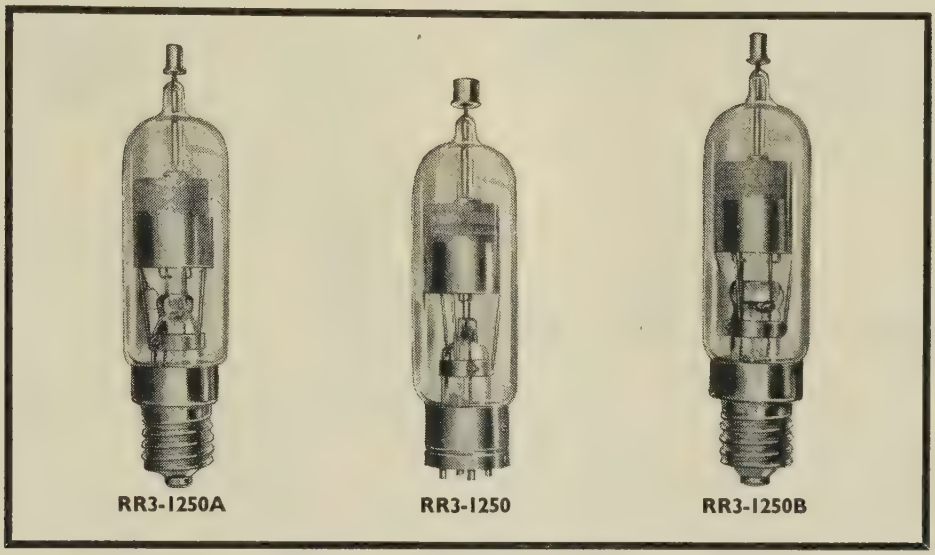
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The Instrument Company of the Cossor Group

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Telephone: CANonbury 1234 (33 lines) Telegrams: Cossor, Norphone, London Cables: Cossor, London

3 new high voltage xenon rectifiers



SHORT HEATING-UP TIME · WIDE AMBIENT TEMPERATURE RANGE · NO "CONDITIONING" ON INSTALLATION

Recent developments in the Mullard 1.25A range of xenon rectifiers have led to improved hold-off characteristics and higher reliability. These improvements, together with the already widely recognised advantages of xenon rectifiers, strongly recommend these valves for power supplies in transmitters, r.f. heaters and similar equipments.

Xenon rectifiers can be operated over a wide ambient temperature range, they are not restricted to vertical mounting, they have a short heating-up time and require no "conditioning" on first being put into service. These features make for great operational convenience and the valves are suitable for use in both fixed and mobile equipment under all climatic conditions.

Two of the valves can be used as plug-in replacements for mercury types: the RR3-1250B in place of the RG3-1250; and the RR3-1250A in place of the RG4-1250 (CV5) in applications where the peak inverse voltage does not exceed 13 kV. Write on your company notepaper to the address below for a free booklet "High Voltage Rectifiers" which gives full data on these and other xenon rectifiers together with details of mercury-filled types.

ABRIDGED DATA

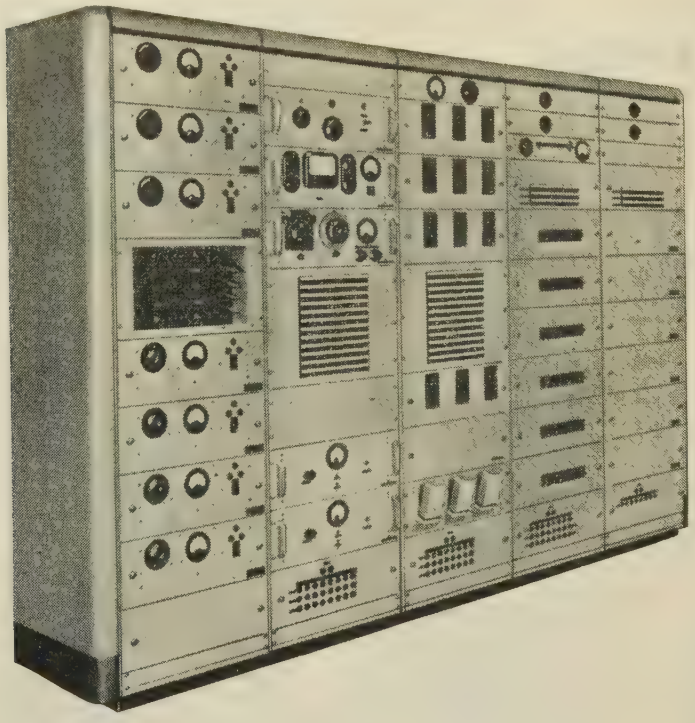
Type No.	Base	Vf (V)	If (A)	P.I.V. max. (kV)	ik (pk) max. (A)	Ik (av) max. (A)	Heating-up Time (secs)
RR3-1250/4B32	B4F	5.0	7.0	10	5.0	1.25	30
RR3-1250A	Goliath Edison Screw	4.0	11.0	13	5.0	1.25	30
RR3-1250B	Goliath Edison Screw	4.0	7.0	13	5.0	1.25	30

Mullard

COMMUNICATIONS AND INDUSTRIAL VALVE DEPARTMENT

MULLARD LTD · MULLARD HOUSE · TORRINGTON PLACE · LONDON W.C.1

MVT 330





telegraph speeds of 280 bauds

The overall performance of long distance radio link telegraph systems can, in many cases, be greatly improved by the introduction of Siemens Ediswan Frequency Modulated Voice Frequency Telegraph Equipment on the land line circuits feeding remotely located radio transmitters or receivers. This equipment provides six channels, five of which will each handle telegraph traffic at speeds up to 280 bauds and one for medium speed traffic up to 140 bauds.

Developed by Cable & Wireless, this telegraph system is now in extensive use by the British Post Office and various overseas administrations.

The equipment is completely self contained and includes full testing facilities for lining up and routine maintenance. The equipment practice employed facilitates removal of individual units. As an alternative, equipment can be supplied which provides 12 channels for handling telegraph traffic at speeds of up to 140 bauds.

The F.M.V.F.T. equipment may be used on any 4-wire circuit capable of transmitting the band 300 c/s to 3400 c/s.

Further information will gladly be sent on request.

extending → **the frontiers of telecommunications**

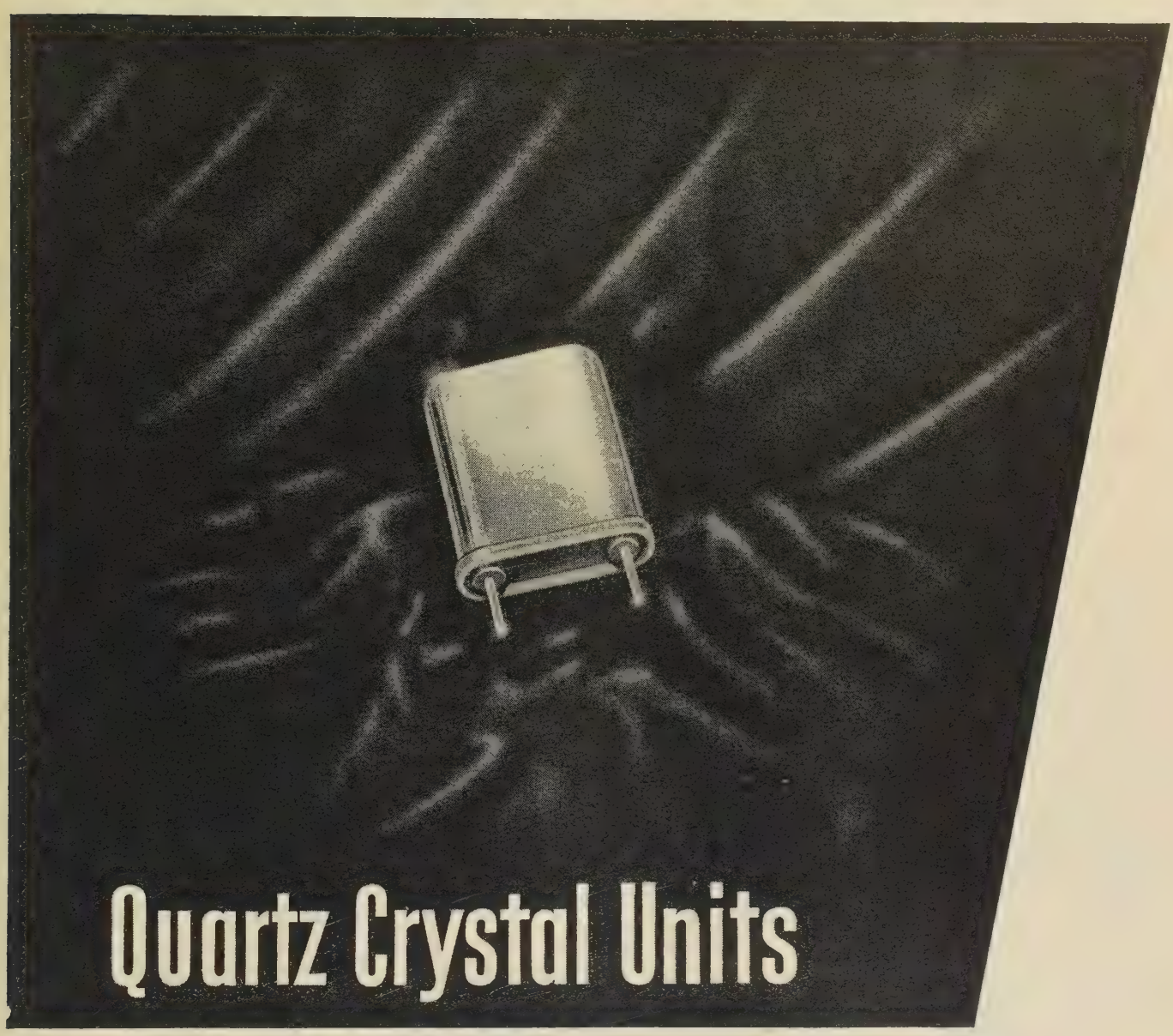
SIEMENS EDISWAN

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Telecommunications Transmission Division, Woolwich, London SE18, England

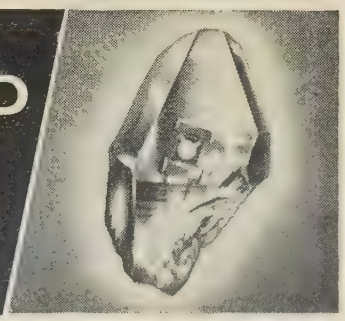
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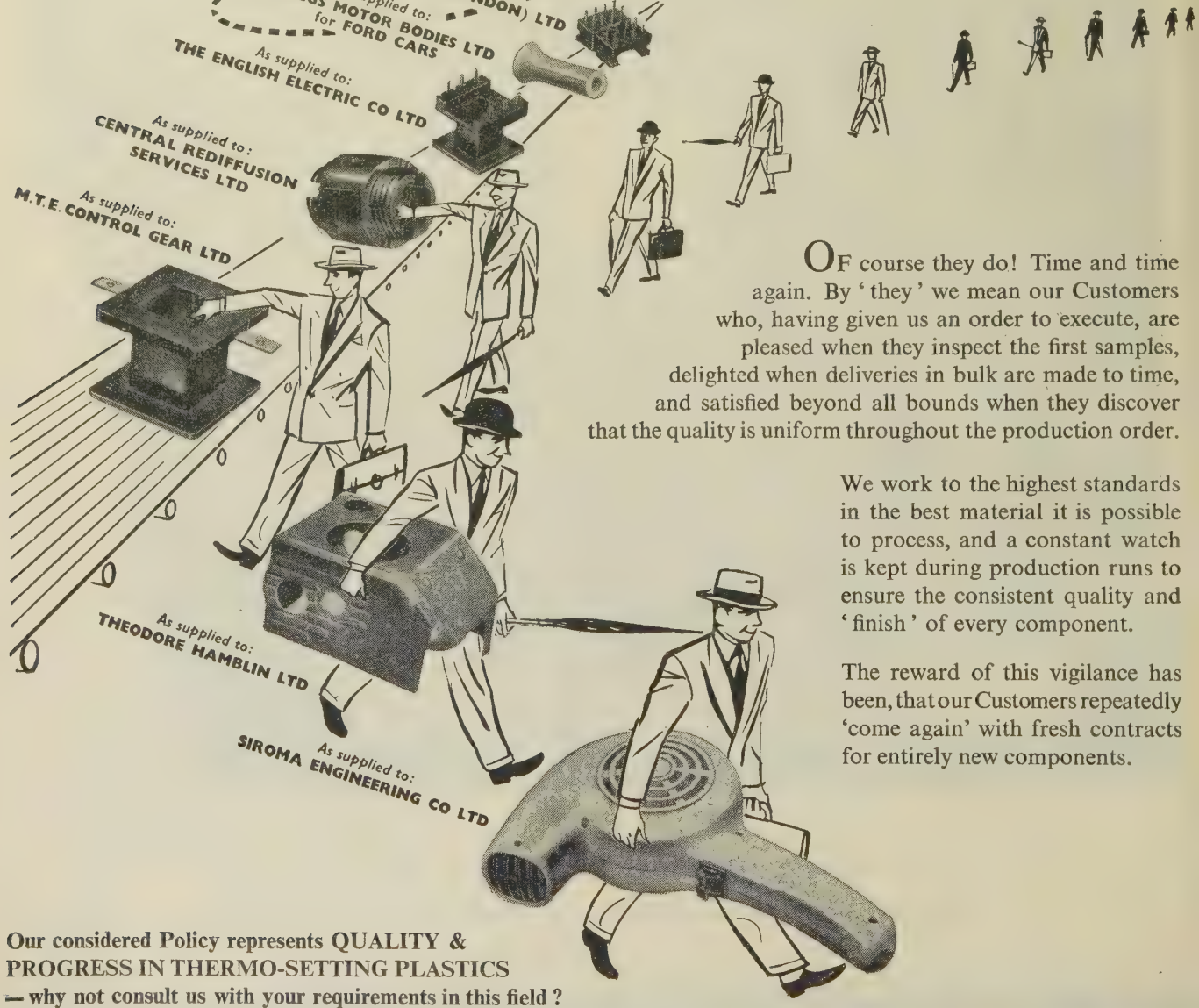
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OF course they do! Time and time again. By 'they' we mean our Customers who, having given us an order to execute, are pleased when they inspect the first samples, delighted when deliveries in bulk are made to time, and satisfied beyond all bounds when they discover that the quality is uniform throughout the production order.

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The reward of this vigilance has been, that our Customers repeatedly 'come again' with fresh contracts for entirely new components.

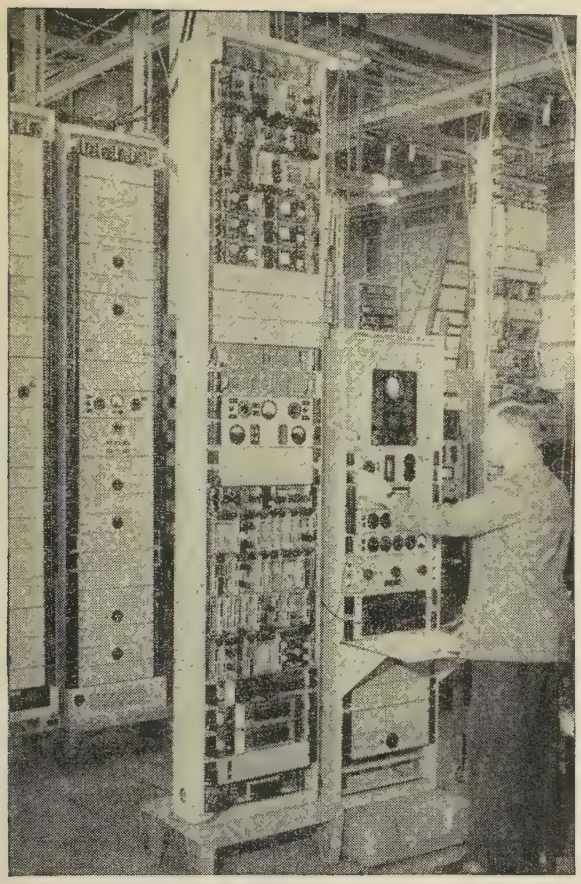
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S.T.C. *developed in 1946 the New Equipment Practice for the unit construction of slide-in panels.*

S.T.C. *developed in 1951 a more compact form of unit construction which has been accepted as the model for up-to-date equipment.*

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S.T.C. *developed in 1955 the new channel panel which has been adopted by leading organisations as the model for their present equipment.*

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Transistorised 12-channel, 2-wire, carrier-on-cable telephone system.

A transistorised 3-channel, open-wire, carrier telephone system.

A 7000 Mc/s portable television outside-broadcast system.

A transistorised 24-channel, frequency modulated, voice-frequency telegraph system.

A 12 Mc/s coaxial system handling up to 2700 high quality telephone circuits or 960 telephone circuits and a television channel.

A transistorised 10-channel rural carrier system.

A small-diameter coaxial cable telephone system for 240 channels, using buried transistorised repeaters.



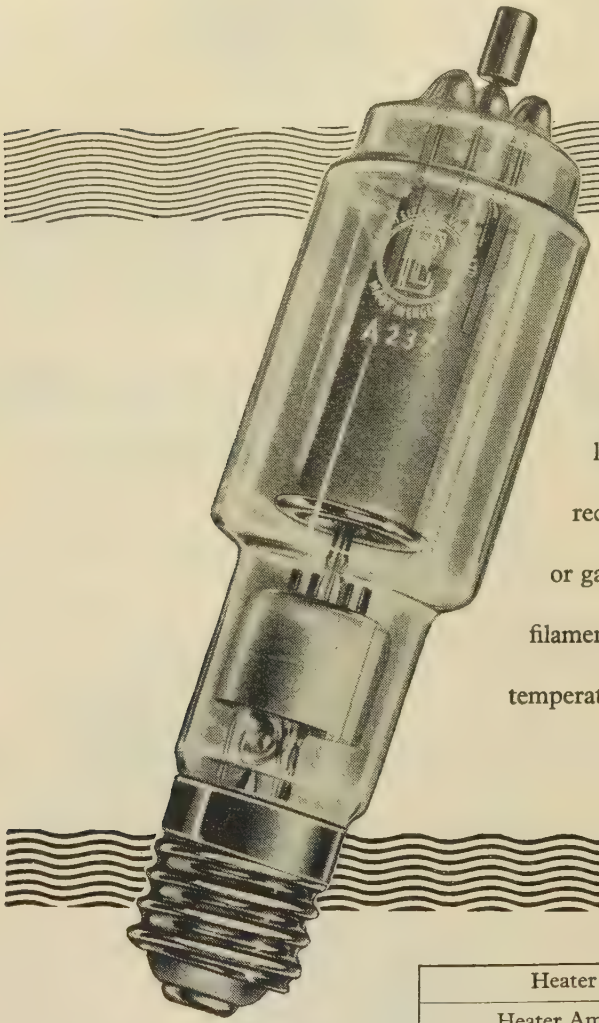
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THE VALVE FOR HIGH VOLTAGE D.C. POWER



The new A237 high vacuum rectifier is one of a range of valves designed for use in laboratory, research and industry where voltage requirements cannot be satisfied by use of mercury or gas-filled rectifiers. It has a thoriated tungsten filament providing stable operation, independent of temperature and with a low voltage drop.

Heater Volts	4
Heater Amperes	12
Peak Inverse Voltage	65 kV
Peak Anode Current	1.5A
Length	235 mm (9¼″)
Diameter	60 mm (2⅜″)
Base	Goliath Edison Screw

Typical operational values in various types of circuit with choke input are as follows :

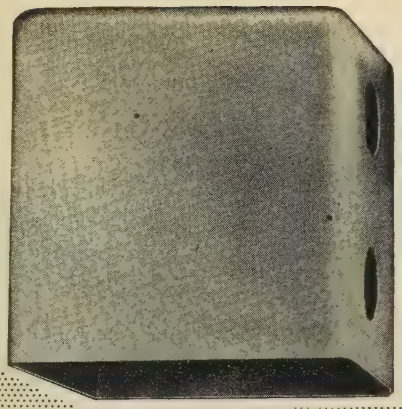
	Single Phase Bridge	Double Three Phase	Three Phase Full-wave
D.C. Current	0.44A	0.90A	0.60A
D.C. Voltage	40 kV	30 kV	60 kV
D.C. Load Output	17.5 kW	27 kW	36 kW

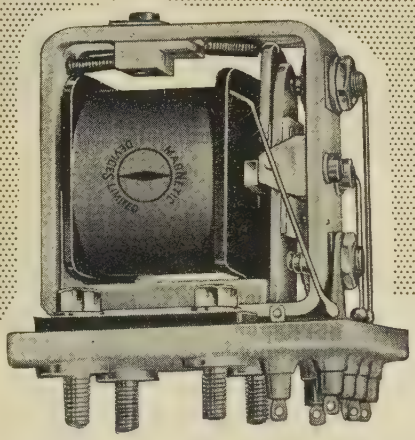
'ENGLISH ELECTRIC'

ENGLISH ELECTRIC VALVE CO. LTD. E *Chelmsford, England*
Telephone: Chelmsford 3491

AP/83

Voltage Regulating Relay





25·5

ZA44704
ZA44706

23·5

12·75

ZA44705
ZA44707

11·75

Two forms of the relay are available, either fully hermetically sealed or enclosed and tropicalised but unsealed. The inter-service reference numbers are as follows:—

Unsealed

ZA 44706 25.5V. make 23.5V. break
ZA 44707 12.75V. make 11.75V. break

Sealed

ZA 44704 25.5V. make 23.5V. break
ZA 44705 12.75V. make 11.75V. break

The Voltage Regulating Relay was designed in co-operation with S.R.D.E., to reduce voltage variations in certain essential circuits of radio sets and has many other applications of a similar nature.

This is particularly necessary in the case of vehicle-borne equipment, with power supplies consisting of lead acid batteries and a small charging generator.

The armature is balanced to withstand vibration, and the complete relay has been subjected to severe vibration testing.

Magnetic shielding is achieved by the iron case, enabling the relay to be used within reasonable proximity of transformers, chokes, etc.

The Voltage Regulating Relay complies with the stringent Ministry of Supply specification No. 166/1, to operate within tolerance, over a temperature range of − 40°C. to +85°C.

MAGNETIC DEVICES LIMITED

MAGNETIC DEVICES LIMITED

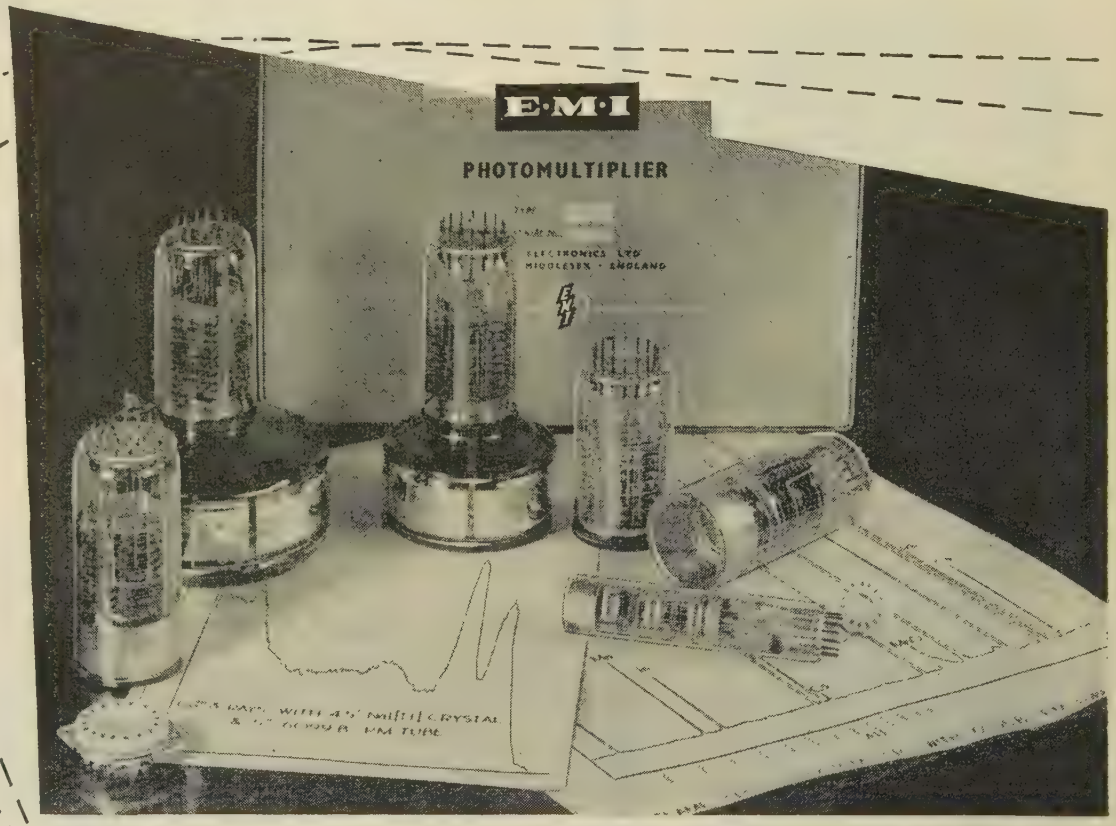
A.I.D. & A.R.B. approved

EXNING ROAD, NEWMARKET, SUFFOLK

Telephone: Newmarket 3181/2/3 Telegrams: Magnetic Newmarket

MD 20

E·M·I Photomultipliers



E.M.I. Electronics manufacture a wide range of end-window photomultiplier tubes having 11 and 13 dynodes giving gain factors up to 5 x 10^8. Tubes are available with glass windows $1\frac{1}{8}$″, 2″, $3\frac{1}{2}$″ and 5″ in diameter and with quartz windows $1\frac{1}{8}$″ and 2″ in diameter. Blue sensitive CsSb photocathodes, of either S11 or a specially made low red sensitivity type, are available in all tube sizes; red sensitive Bi-Ag-Cs photocathodes (S10) are available with glass and quartz windows in the $1\frac{1}{8}$″ and 2″ sizes.

Suggested types for specific applications are:—

	APPLICATION	TYPE
SCINTILLATION COUNTING	Gamma ray spectrometers and general counting	9524, 6097, 9514, 9531, 9530, 6099.
	Counting of low energy particles (C^{14}, H^3).	9524S, 6097S, 9514S (low red cathode).
	Observation of gas scintillations.	6255, 6256 (quartz windows).
SPECTROPHOTOMETERS & ASTRONOMICAL APPLICATIONS	3500—6500 A.U.	9502, 6097, 9514, 9524.
	1600—6500 A.U.	6255, 6256.
	3500—8500 A.U.	6095, 9528.
	1600—8500 A.U.	9526 ($1\frac{1}{8}$″ quartz window, Bi-Ag-Cs).
LOW LEVEL PHOTOMETERS	Corrected to eye response by filter.	6095, 9528.
FLYING SPOT SCANNERS FOR TELEVISION	Monochrome	6097F, 9530F.
	Colour	9530F, 6097F, 6095F.

For data sheets and descriptive booklet giving full details of E.M.I. photomultipliers and other types of electronic tubes and further recommendations for particular applications, apply to:—

E.M.I. ELECTRONICS LTD., (Valve Division) **RUISLIP, MIDDLESEX, ENGLAND**

Cables: **EMIDATA, RUISLIP, ENGLAND**

EE63

When it's a question of Flexibility...

the SERVOMEX L.F.51 low frequency wave-form generator leads the World!

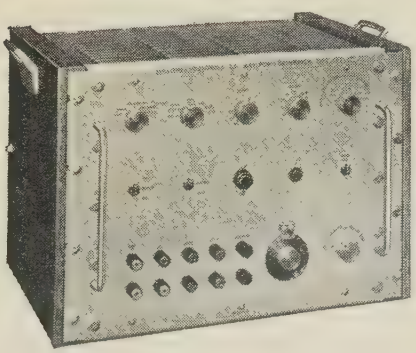


The L.F.51 is an all-British Function Generator of patented design giving a flexibility that has not even been approached by any other instrument. It has now been in production for nearly two years and has been widely adopted for driving analogues and real systems in the United Kingdom and abroad. (Exports, including U.S.A., are over 25% of sales.)

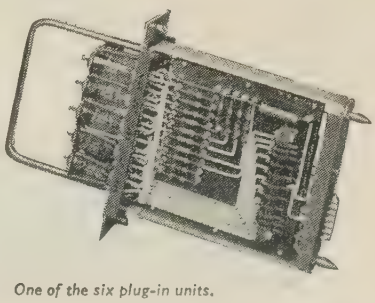
37 Different Waveforms can be generated

SINEWAVES (500C/s down to 1 cycle every 33 minutes)
SQUARE WAVES AND PULSES 100μS to 1,000 secs (rise time 5μS)
RAMPS (lasting 1 millisecond to 1,000 seconds)
Single or repeated pulses of square, triangular, sawtooth, cosine, trapezium shape, sine squared, etc. With modification of 1 unit, a variety of non-standard shapes can be simulated in either single transitions or pulses, e.g. Gaussian.
VOLTAGE 150 volts to less than 100 microvolts peak to peak
LOAD current up to 5mA peak
4 internal stabilised supplies, to maintain frequency and amplitude calibration
Plug-in construction for ease of servicing and compactness
Special synchronising circuit to trigger CRO etc., in advance of output wave
Decade frequency setting. Balanced (reversing) output

Technical Data Sheets available on request.



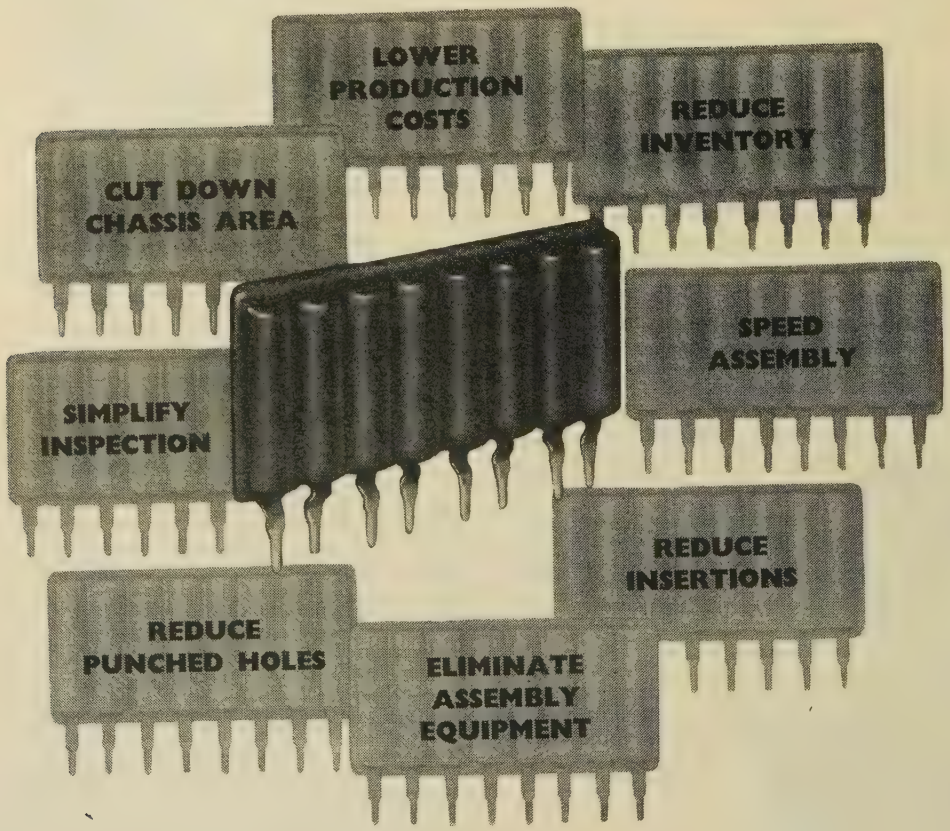
The L.F.51 with wooden ends, removable for use in the 19in. rack.



One of the six plug-in units.

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The Erie pre-assembled component system, known as "Pac", is a thin vertical module containing standard resistors and capacitors, of proven quality, mounted to a printed wiring board in a stable mechanical assembly, tailored to the requirements of the individual customer.

Besides offering considerable savings in itself, "Pac" paves the way for fully three-dimensional assembly, thus enabling the designer to take the fullest possible advantage of the potential savings in space attendant upon the introduction of shorter, wide-angle, tubes for television.

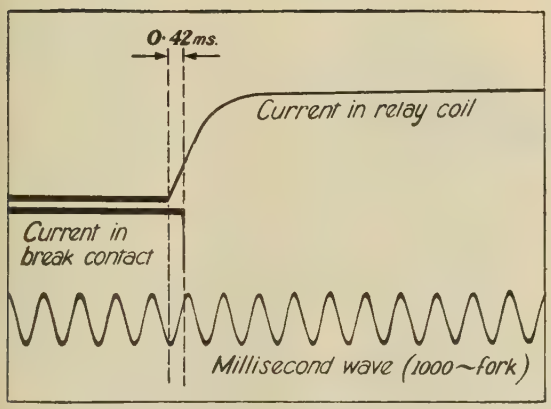
ERIE★ Resistor Ltd

★ Registered Trade Mark

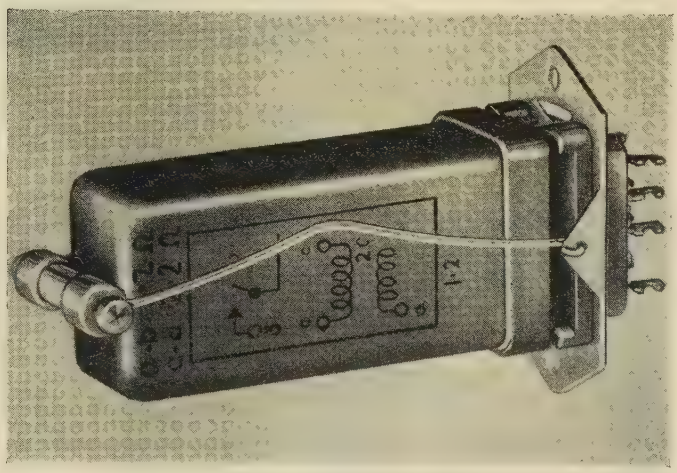
Carlisle Road, The Hyde, London, N.W.9., England. Tel: COLindale 8011. *Factories: London and Gt. Yarmouth, England; Toronto, Canada; Erie, Pa., and Holly Springs, Miss., U.S.A.*

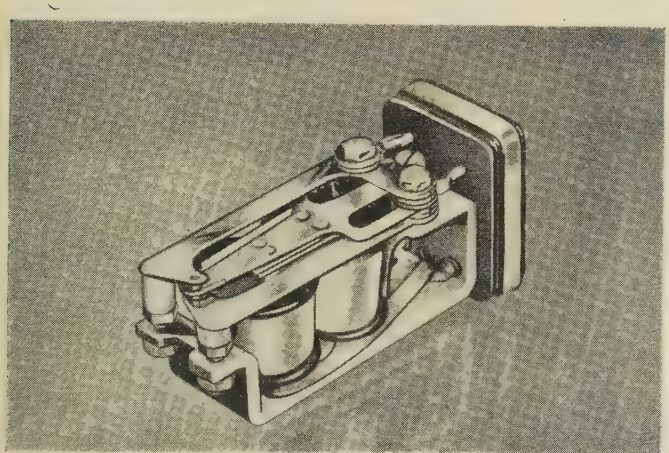
SIEMENS EDISWAN

Miniature *high speed* relay



Oscillogram of the operating-lag of the break contact





This small relay has earned for itself a world-wide reputation on account of its very rapid operation, great reliability and insensitivity to external mechanical and electrical disturbances. It was originally designed to meet very severe conditions called for by the Services; it is now in use in vast numbers for all manner of applications and varying conditions.

Inexpensive and available for early delivery.

Can be supplied with plug-in base if required to form a readily interchangeable plug-in unit.

Hermetically sealed, unaffected by dirt or immersion in water and immune to wide or rapid changes of temperature or air pressure.

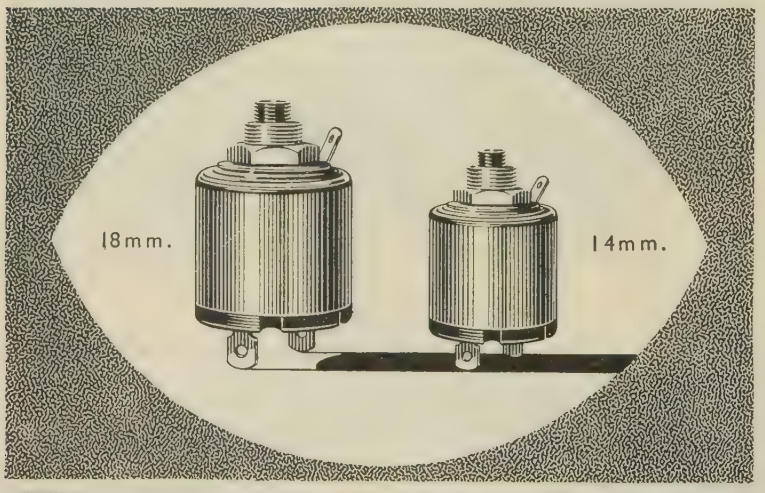
We shall be pleased to give full details if you telephone (Woolwich 2020, Extension 621) or write to:

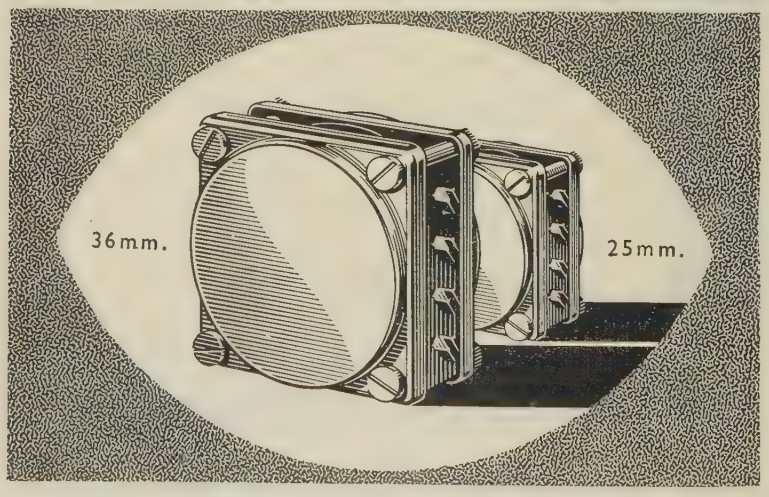
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* Operative over a wide frequency range
* Controlled temperature coefficient

Wherever high quality pot cores are required, there will be a Mullard type available to meet the specification, furthermore, they can be supplied wound to customers individual requirements.

Write now for full details of the comprehensive range currently available.

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MULLARD LTD., COMPONENT DIVISION, MULLARD HOUSE, TORRINGTON PLACE, W.C.1

MC 255A

"Why work them out by hand?"

"How else?"

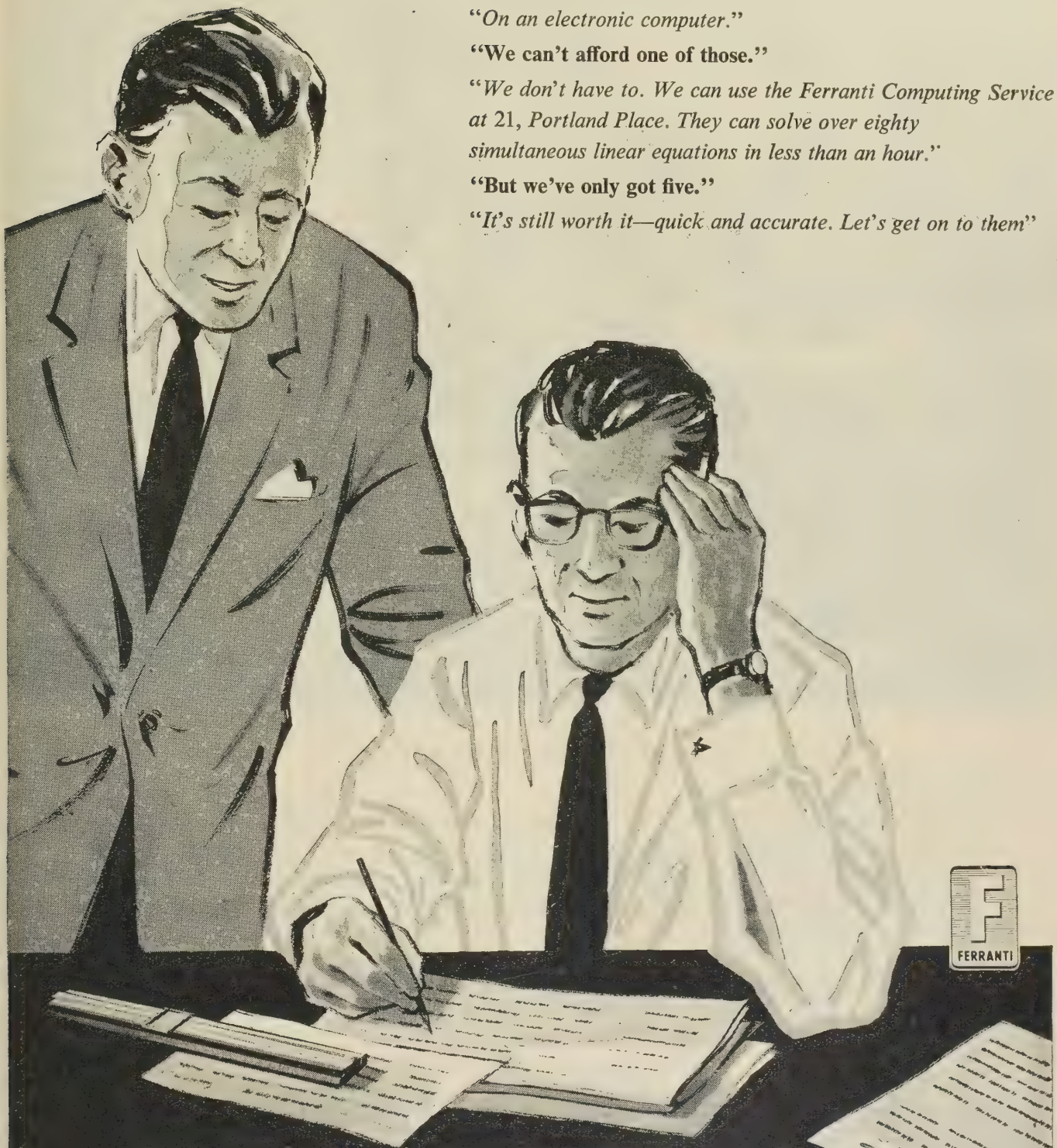
"On an electronic computer."

"We can't afford one of those."

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"But we've only got five."

"It's still worth it—quick and accurate. Let's get on to them"



For details of the Ferranti Computing Service phone or write to:

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LONDON COMPUTER CENTRE: 21 PORTLAND PLACE, W.1

Telephone: LANgham 9211

Works: **WEST GORTON · MANCHESTER 12**



What's in a name?

As a trade name grows familiar one is apt to forget its derivation. We think that our illustration of a core being assembled for a Savage Massicore Transformer may serve as a reminder that all our instruments are constructed with massive cores. Such stout-heartedness is essential for the faultlessly high performance expected from Savage Transformers (rightly so) by electrical and electronic engineers everywhere.

It is, however, only part of the story. Savage Massicore Transformers owe their world-wide good name to the critical choice of allied materials and the meticulous skill with which they are combined into a solid and reliable whole.

MASSICORE SAVAGE

SAVAGE TRANSFORMERS LTD

NURSTEED ROAD · DEVIZES · WILTSHIRE · Tel: Devizes 932

Corner for Contented Customers

"Gentlemen:

By good authority we have been told that you make some of the best transformers ever made . . .

H. E. California."

TP/56

world-wide approval

Pye Telecommunications Limited are now marketing the widest and most modern range of V.H.F. fixed and mobile radio-telephone equipment available in the world. This range of equipment has been designed to expand the application of Pye Radio-Telephones already in constant use all over the world.

Pye Ranger V.H.F. equipment has now received approval from the British G.P.O. for Land, Marine and International Marine applications employing A.M. or F.M. systems, type approval from the Canadian D.O.T., and type acceptance of the F.C.C. of the United States of America.

No other Company holds so many approvals for this range of equipment, which now covers every conceivable requirement.

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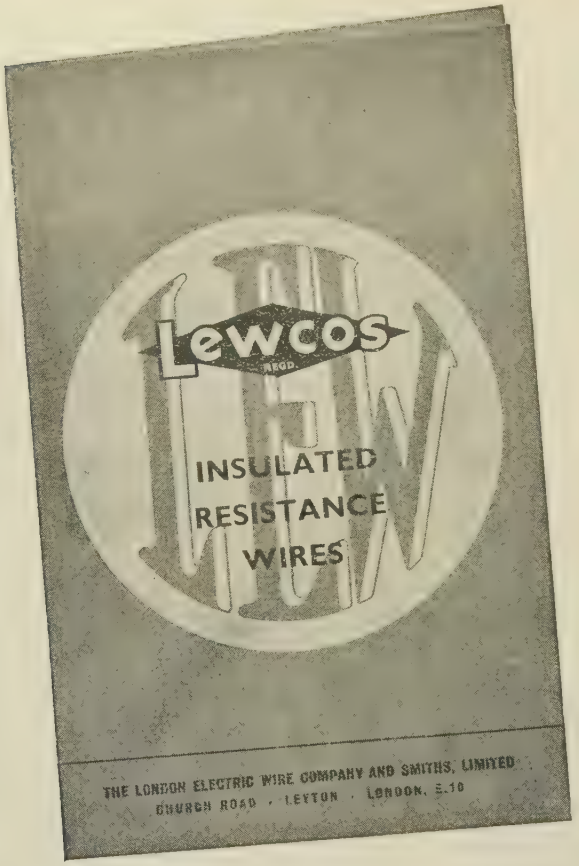
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Felgate Electronics Limited announce a new range of Wire Wound Precision Resistors which offer utmost reliability, prompt delivery, and exceptionally high stability (0·02 %). The specification speaks for itself:

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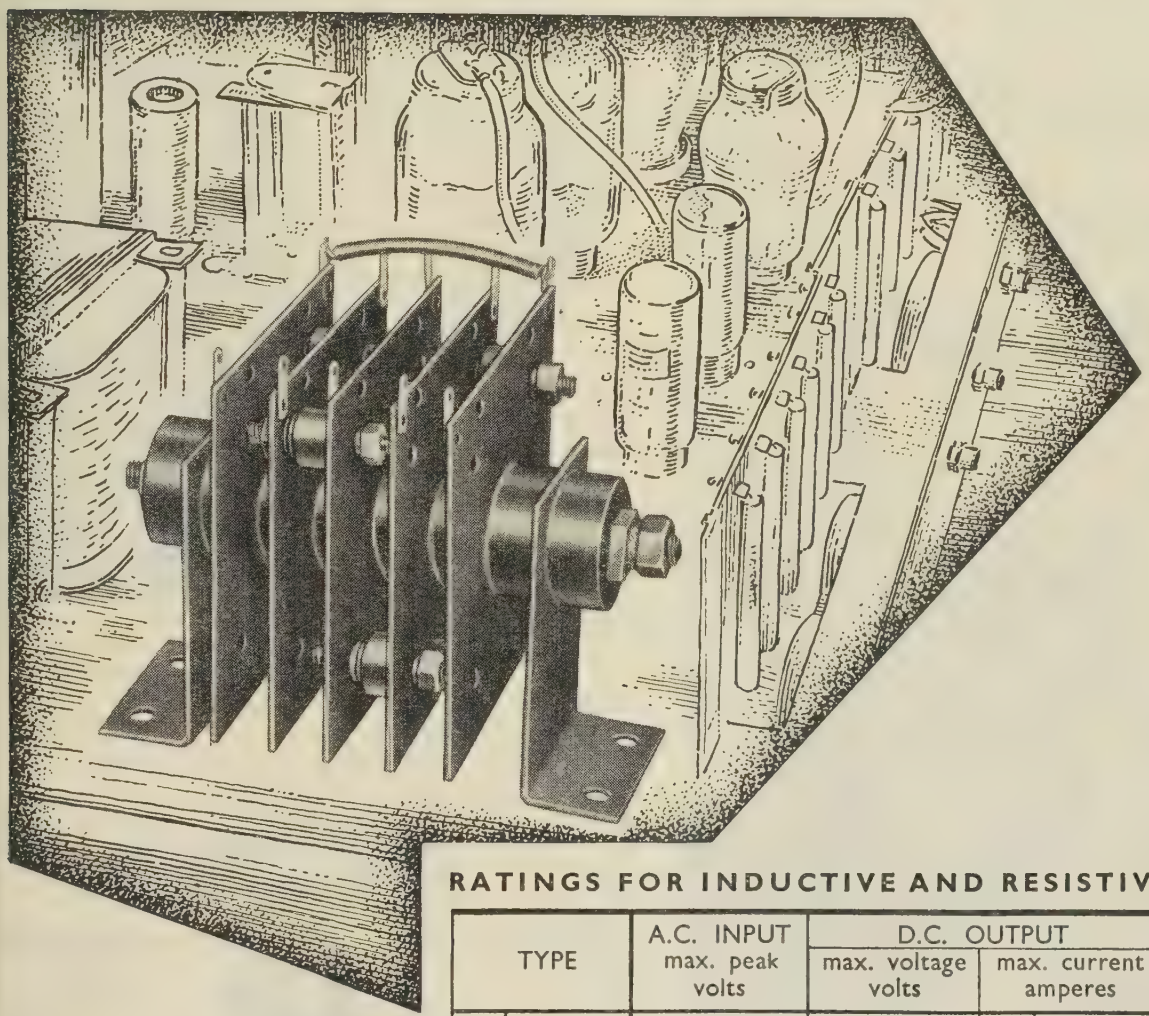
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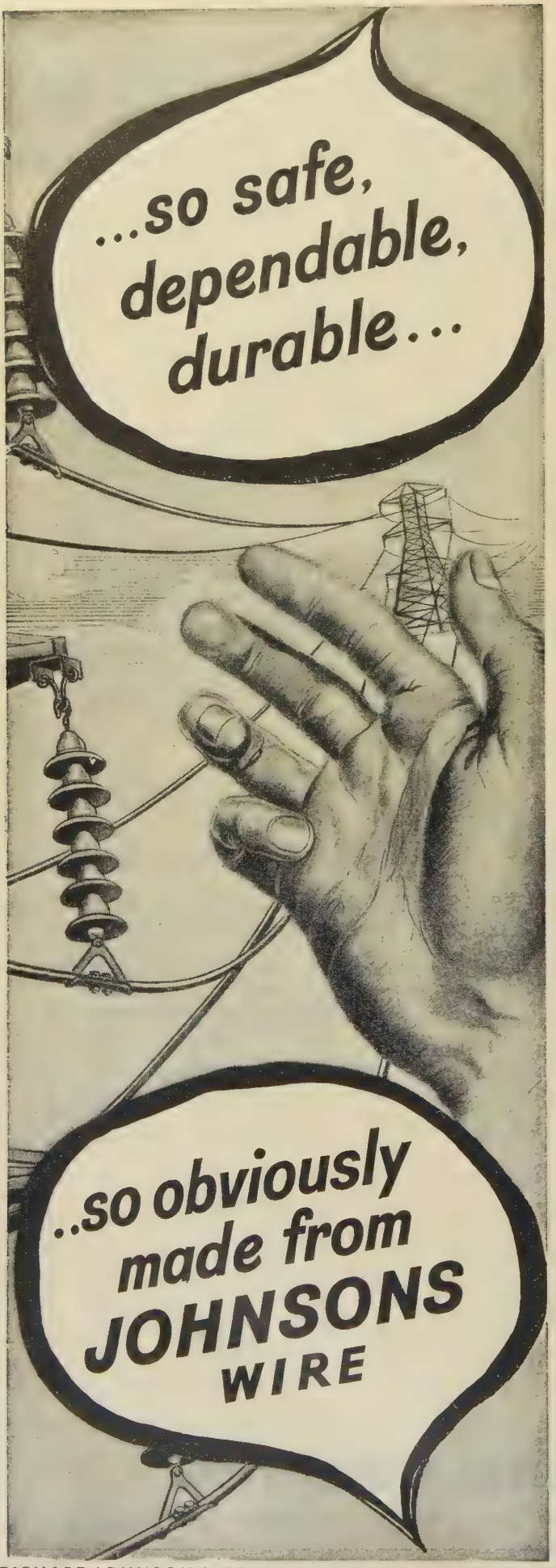
	TYPE	A.C. INPUT max. peak volts	D.C. OUTPUT max. voltage volts	D.C. OUTPUT max. current amperes	OVERALL LENGTH inches
SINGLE-PHASE	GA31-A	140	125	2·0 at 40°C	$3\frac{1}{2}$
	GA41-A	53	47	2·0 at 60°C	$3\frac{1}{2}$
	GA51-A	210	187	2·0 at 40°C	$3\frac{1}{2}$
	GA61-A	106	94	2·0 at 60°C	$3\frac{1}{2}$
	GA52-A	340	303	2·0 at 40°C	5
	GA62-A	170	151	2·0 at 60°C	5
	GA53-A	510	455	2·0 at 40°C	$6\frac{1}{2}$
	GA63-A	254	227	2·0 at 60°C	$6\frac{1}{2}$
THREE-PHASE	GB31-A	140	188	3·0 at 35°C	$4\frac{5}{8}$
	GB41-A	53	71	3·0 at 55°C	$4\frac{5}{8}$
	GB51-A	210	283	3·0 at 35°C	$4\frac{5}{8}$
	GB61-A	106	143	3·0 at 55°C	$4\frac{5}{8}$
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	GB62-A	170	229	3·0 at 55°C	$6\frac{3}{4}$

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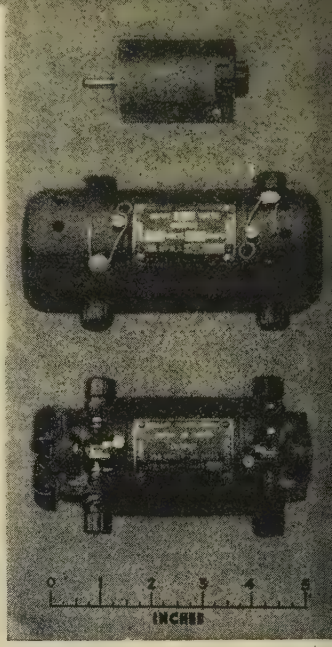
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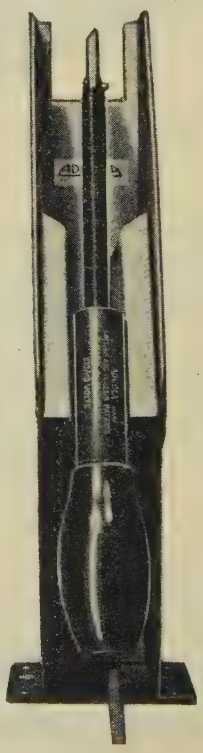
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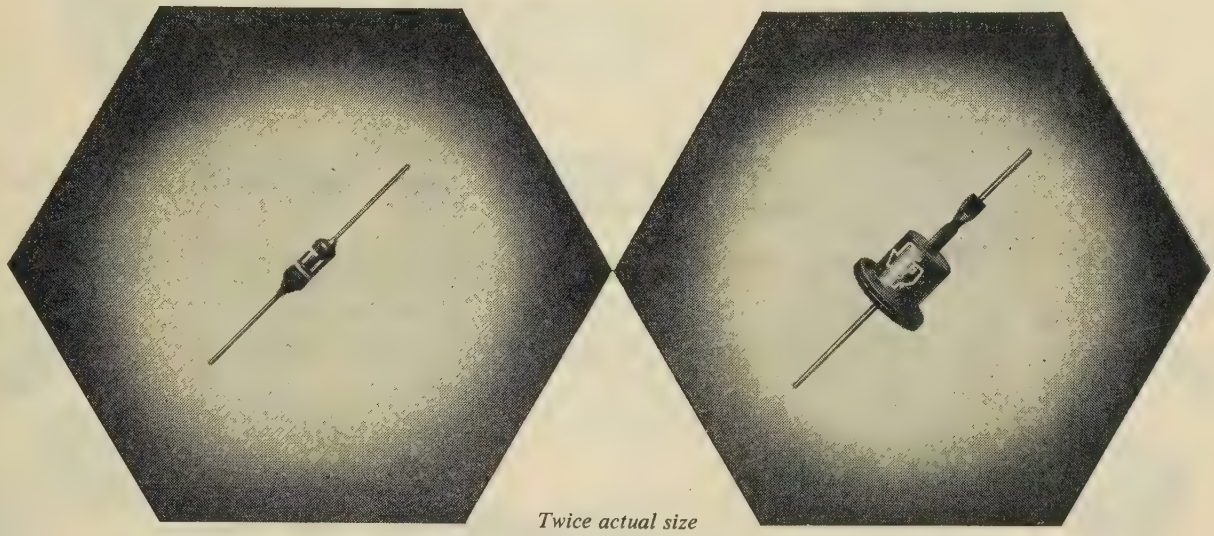
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				'MOLY G'	'METAL CASE'		
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	Average rectified Forward Current at 150°C			150	250	mA	
	Recurrent Peak Forward Current at 25°C			1.25	2.5	Amp.	
	Surge Current 1 Sec. D.C. at +25°C to 150°C			3.0	15.0	Amp.	
SPECIFICATIONS	Max. Reverse Current at P.I.V. 25°C			0.2	10.0	μA	
	Max. Reverse Current at P.I.V. 100°C			15.0	300	μA	
	Max. Voltage Drop at I_o=400 mA at 25°C			1.0	1.0	volt	

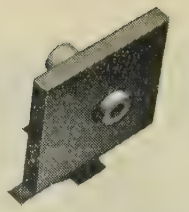
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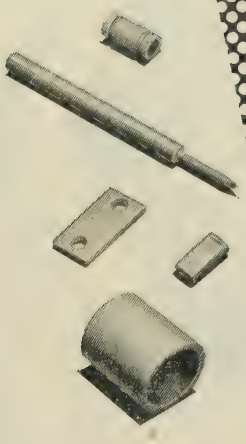
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To illustrate the extreme hardness of SINTOX Industrial Ceramic, the photograph on the right shows a ½ in. diameter tube of SINTOX which, without even chipping, was forced through a ¼ in. thick mild steel plate.



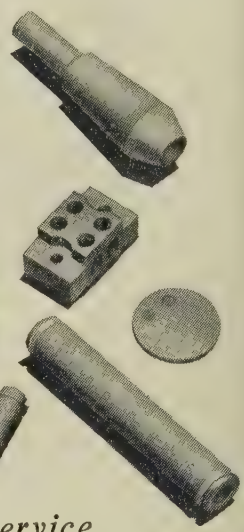
SINTOX IN THE ELECTRICAL INDUSTRY



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Application should be made to

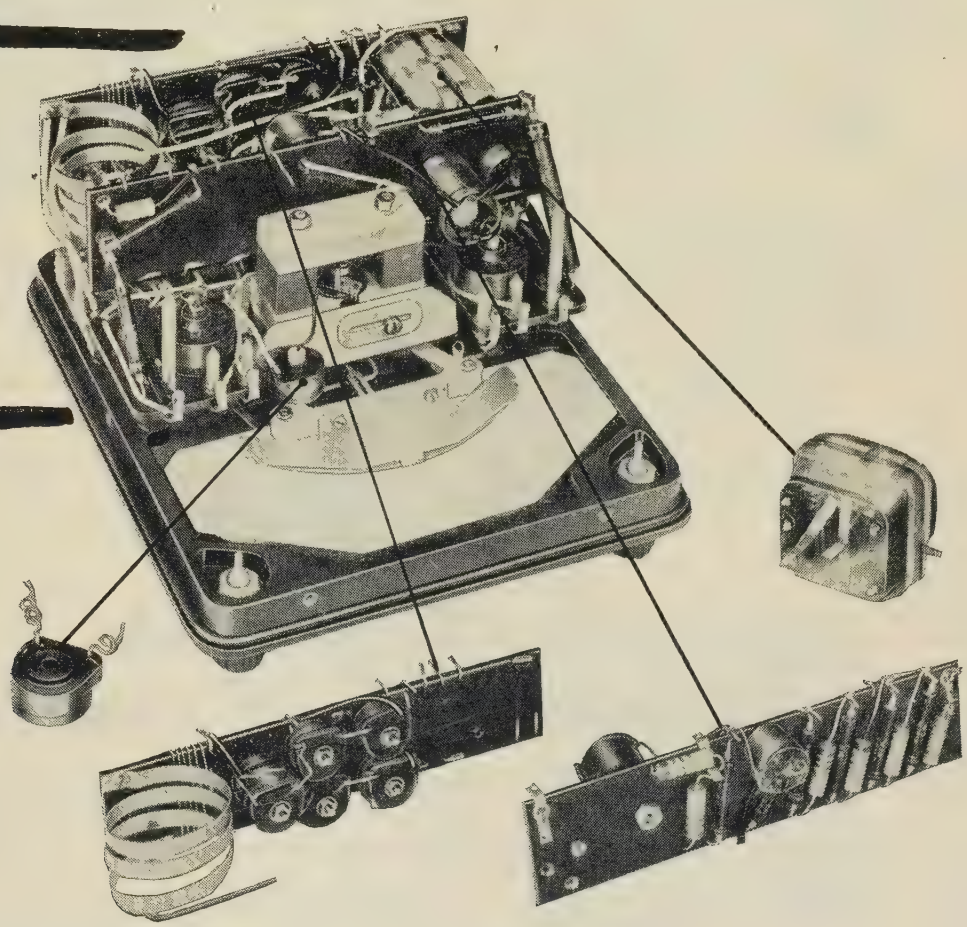
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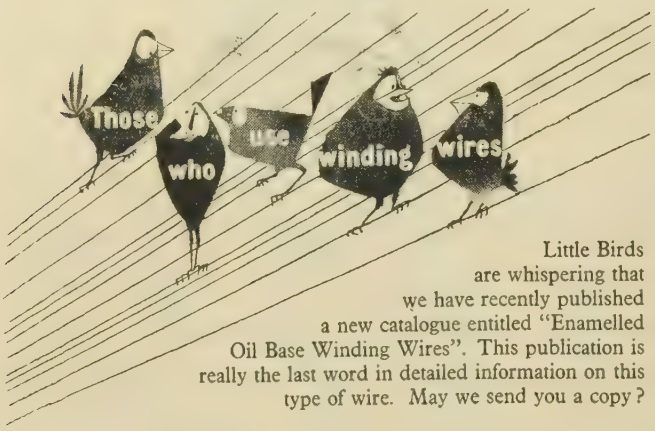
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- ★ **for impregnating, potting or sealing electrical windings and components**
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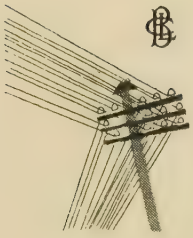
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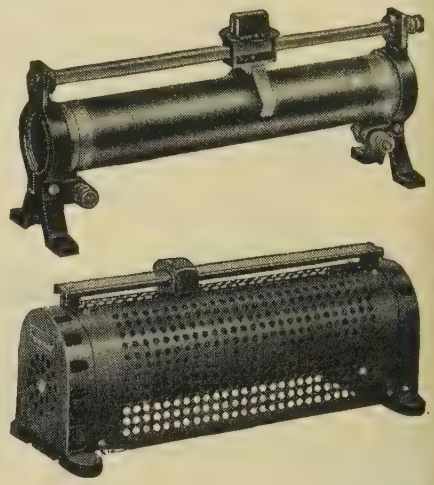
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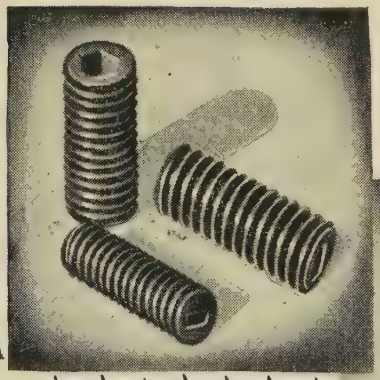
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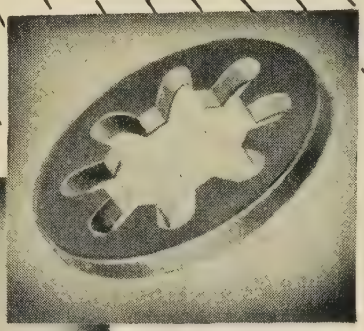
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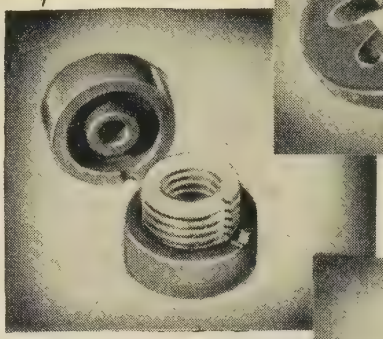
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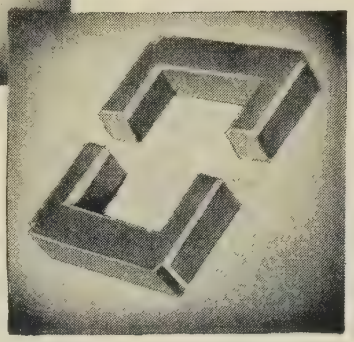




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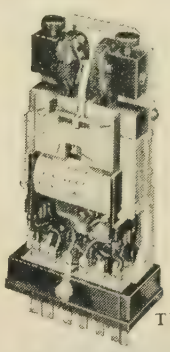
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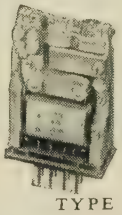
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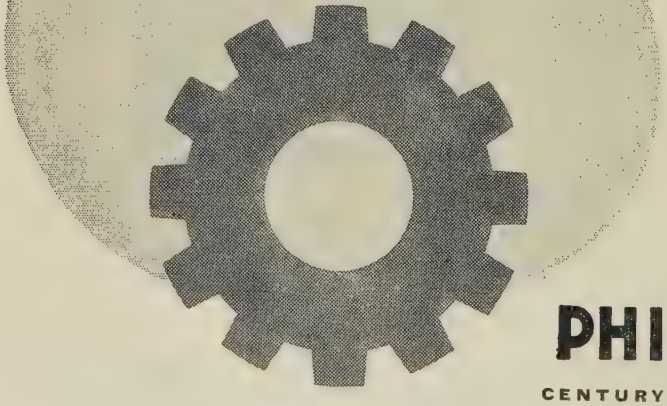
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8μF	12	$\frac{3}{4}$″	$\frac{3}{16}$″
16μF	6	$\frac{3}{4}$″	$\frac{3}{16}$″
24μF	3	$\frac{3}{4}$″	$\frac{3}{16}$″
32μF	1½	$\frac{3}{4}$″	$\frac{3}{16}$″
3μF	12	$\frac{1}{2}$″	$\frac{3}{16}$″
6μF	6	$\frac{1}{2}$″	$\frac{3}{16}$″
8μF	3	$\frac{1}{2}$″	$\frac{3}{16}$″
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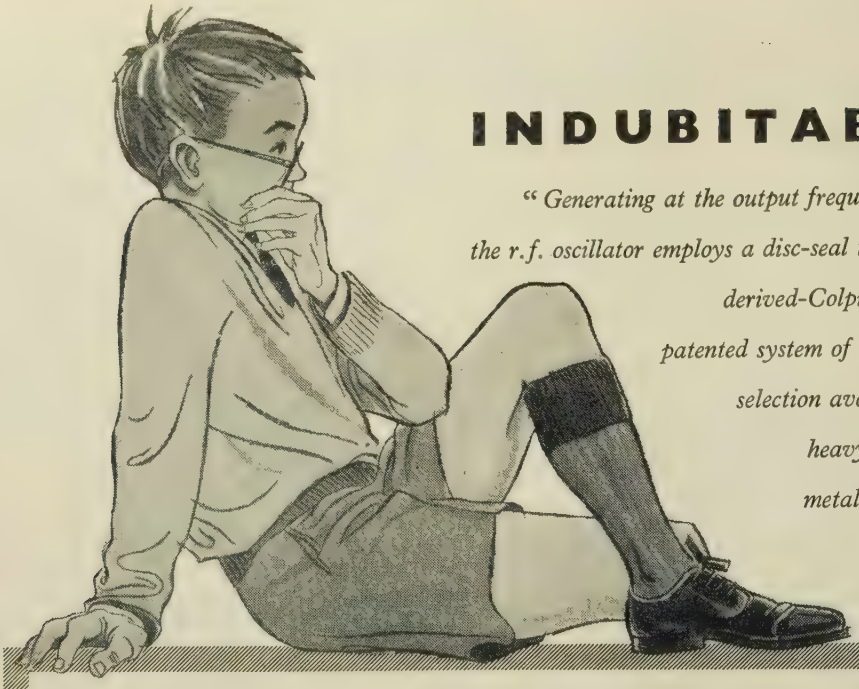
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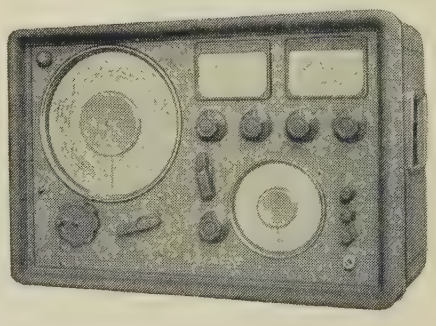


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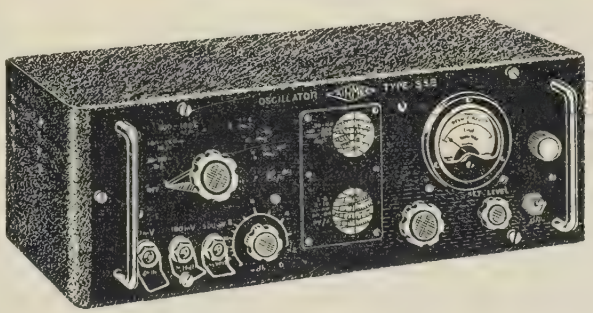
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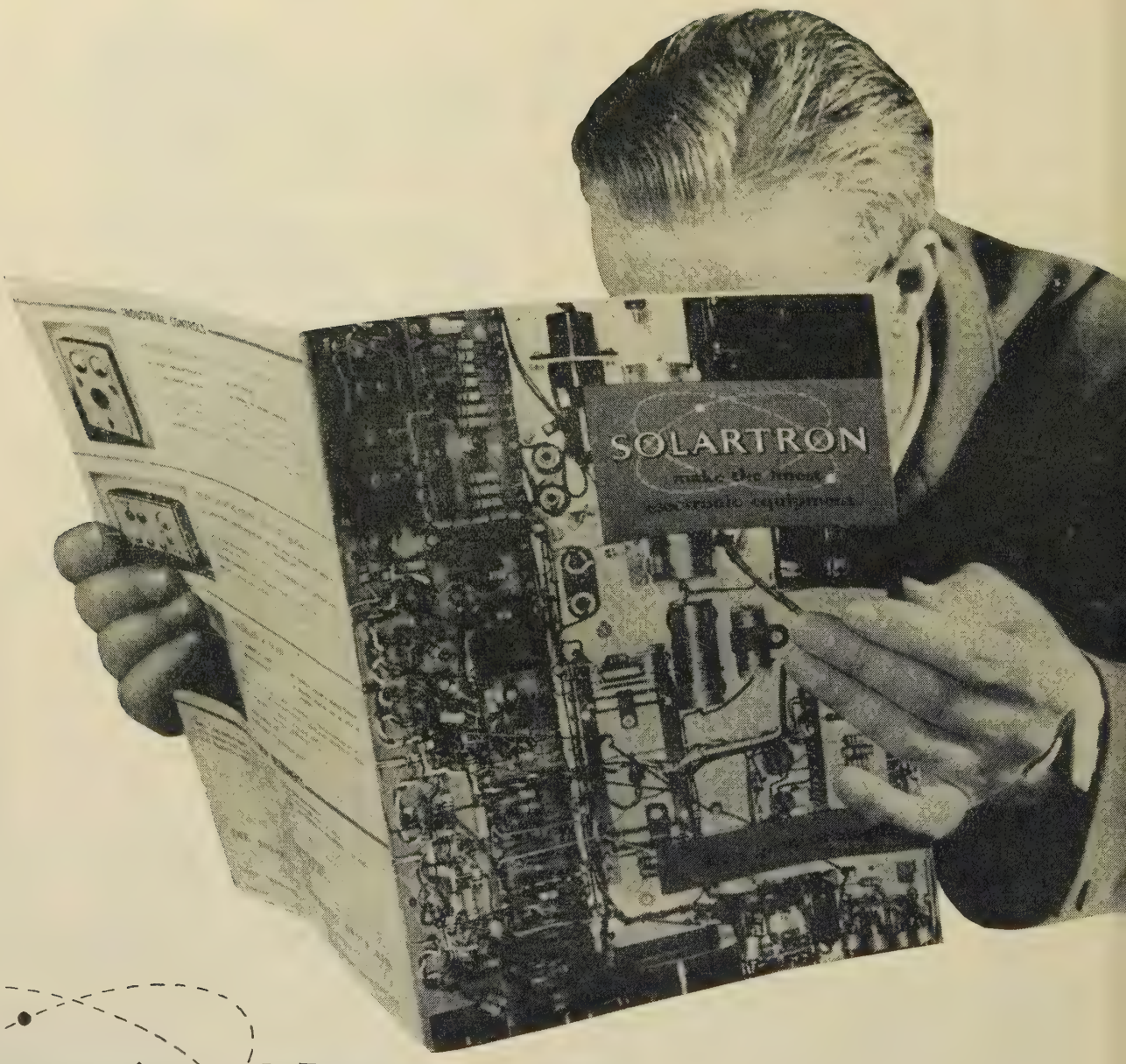
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THE REMOTE AND AUTOMATIC CONTROL OF SEMI-ATTENDED BROADCASTING TRANSMITTERS

By R. T. B. WYNN, C.B.E., M.A., and F. A. PEACHEY, Members.

(*The paper was received 17th October, 1956. It was published in April, 1957, and was read before the* RADIO AND TELECOMMUNICATION SECTION *10th April, 1957*)

SUMMARY

In the Chairman's Address to the Radio Section in October, 1949, a description was given of the steps which the B.B.C. was taking to extend its sound-broadcasting service without a proportional increase in the number of its technical staff. This was being achieved by a process which now might be termed 'automation', but which at that time meant broadcasting from unstaffed transmitters remote from the studios supplying the programme material. The paper reviews this technique and may be regarded as confirmation that these steps were justified.

(1) INTRODUCTION

The Chairman's Address[1] to the Radio Section in October, 1949, discussed the possibility of operating sound-broadcasting transmitters without staff continuously in attendance. The paper describes the progress which has been made, and the lessons which have been learnt, in developing semi-attended stations.

Up to the present, the number of operating personnel which it has been possible to save from the duty of transmitter monitoring, and provide with work of greater interest and promotional opportunity, has been about 200, and there are now 67 B.B.C. m.f. and v.h.f. transmitters operating without continuous supervision by staff. The shut-down times of stations in this group are given in the Appendix, and, for local faults, these may be compared with the shut-down time of 0·019% taken over the same period for fully-attended transmitters.

The design of sound-broadcasting transmitters has reached a point where a very high standard of reliability can be achieved. But, unfortunately, it cannot be assumed that the power supply to the station is never likely to be interrupted, that the life of components, including valves, is becoming infinite, and that the programme sent to the station from some distant point will always arrive without breaks or distortion. For this reason it is necessary for a broadcasting organization to develop equipment which will either report a state of abnormality at the distant transmitting station or itself perform switching operations in emergencies.

The design of this equipment is determined by the importance of the station and also the grade of facilities that can be provided at a reasonable cost. A large and important transmitting station must be critically monitored and most carefully maintained. For this purpose it is desirable to have an engineer available to assess the monitor's indications of trouble and to determine the action which should be taken. A small station, often giving a service supplementary to that from a larger station, requires no engineer in immediate attendance and may be controlled by monitoring gear which takes automatic action only on the serious types of faults. Such factors, together with those of distance and other technical considerations, have necessitated the development of an appropriate range of automatic monitoring and switching equipment. It will be seen that the desire to monitor the performance of line links has been the reason for the development of a large part of this equipment.

(2) DEFINITION OF SEMI-ATTENDED TRANSMITTING STATIONS

In this country there are two main types of semi-attended transmitting station in service, namely those which are remotely controlled from the nearest continuously-staffed centre, and those which are automatically controlled by time switches and electronic equipment. These two types of stations are known as *remotely controlled stations* and *automatically controlled stations*, and these definitions have become accepted by the European Broadcasting Union. The E.B.U. also classifies stations as *unattended* and *semi-attended* in the following way: a *semi-attended* station is not normally manned, but there are staff either on other duties or living nearby. An *unattended* station has no staff either on duty or living nearby on call. The stations described in the paper are therefore classified as *semi-attended.*

In the Appendix, the distance of these stations from the nearest fully-staffed centre, or the centre from which it is convenient to monitor and control, is shown, because it is the factor which largely influences the type of monitoring equipment and determines whether the station should be remotely or automatically controlled.

This is an 'integrating' paper. Members are invited to submit papers in this category, giving the full perspective of the developments leading to the present practice in a particular part of one of the branches of electrical science.

Mr. Wynn and Mr. Peachey are with the British Broadcasting Corporation.

(3) PLANNING REMOTELY-CONTROLLED AND AUTOMATICALLY-CONTROLLED STATIONS

It has been realized from the outset that transmitters in this category must be designed for the best inherent reliability, which should not be attained by adding supervisory equipment, for this might itself develop a fault and close down the transmitter unnecessarily. Here, the term *transmitter* includes the valves and all the transmitter auxiliaries. This is the simple policy of prevention being better than cure. Even so, special electronic equipment has had to be developed to perform in emergencies some of the functions normally carried out by the shift-keeping staff at attended stations. The desired reliability has been attained by operating a number of unit transmitters, with their inputs in parallel and their outputs connected, through a combining unit, into a common aerial. The failure of one unit therefore causes a reduction in, but not the cessation of, the radiated output power. Transmitters of this type are referred to in the paper as *paralleled transmitters*.

To a large extent it has been possible to standardize the auxiliary automatic equipment used for these two types of station. Variations in the design of equipment, however, are brought about by the distance over which monitoring apparatus must operate and the extent to which the station can be visited. For instance, where a complete programme signal loop can be provided either by radio reception or by a return line from the transmitter, the simplest monitor application is possible, and the *automatic monitor minor*[2] was developed for this purpose. On the other hand, the monitoring loop must be completed by other means if the distance is so great that radio reception is out of the question and the cost of a return line is excessive. For this purpose the Post Office have co-operated by providing the earth-phantom circuit derived from the programme line. Although this channel is suitable for conveying the required information concerning the programme signal, it is so restricted in frequency range that the minor is unsuitable and alternative apparatus has had to be developed. There are two types of monitor which use such channels, namely the *executive monitor* and the *extended minor*.

Executive apparatus is so called because it is designed both to discover and to clear a fault. The extended minor operates in a similar way to the original minor in the sense that it gives an alarm only when a fault arises, and it depends on the discretion of technical staff who decide whether it is expedient to clear the fault immediately. The executive apparatus, lacking this discretion, must be subject to restraint, and must be allowed to take action only in necessity; for this purpose the *line executive monitor* has been designed.[3]

Since these unattended stations employ several transmitter units in parallel, it has been simpler with the executive equipment to divide the monitoring into two parts. Thus, in addition to the line executive monitor, there is a *transmitter executive monitor* which is suitable for keeping a watch on each of the individual transmitter units. These several designs of monitor have been employed to cover all the B.B.C.'s monitoring requirements. Variations in grade of performance, etc., are obtained by using auxiliary equipment. For example, if executive monitors cause a station to shut down, special signalling facilities are provided to enable staff at the nearest manned station to diagnose the trouble and, where possible, restore the service.

(4) DESCRIPTION OF MONITORING EQUIPMENT

(4.1) The Automatic Monitor Minor

A description of this monitor and the various ways in which it can be applied has been given before[2] and need not be repeated here. Suffice it to say that this monitor has been applied extensively to both the automatically and remotely controlled stations. It is primarily used for monitoring where a closed loop by line or radio can be provided for the programme signal. The distance over which this monitor can be operated is conditioned by a number of circumstances, but in general it can be applied if the unattended station is within about 30 miles of the parent station. Also, since this monitor operates to a very critical standard, it is dependent on technical staff being available to assess the action that should be taken when it alarms. It therefore has certain limitations in its application.

(4.2) Line Executive Monitor

As explained earlier, this equipment makes use of the earth-phantom channel on the programme line. This phantom circuit, although restricted in its transmission characteristics, is adequate for sending information describing the broad parameters of the programme signal.

The programme signal at the point of origin is rectified and 'gated' to provide three distinct conditions corresponding to high, medium and low level programme. The time-constants of the circuits are such that these signals may be used to operate telegraph-type relays which in their three equivalent positions send different d.c. signals to the distant transmitting station. The nature of the signals is shown in Fig. 1. These are transmitted

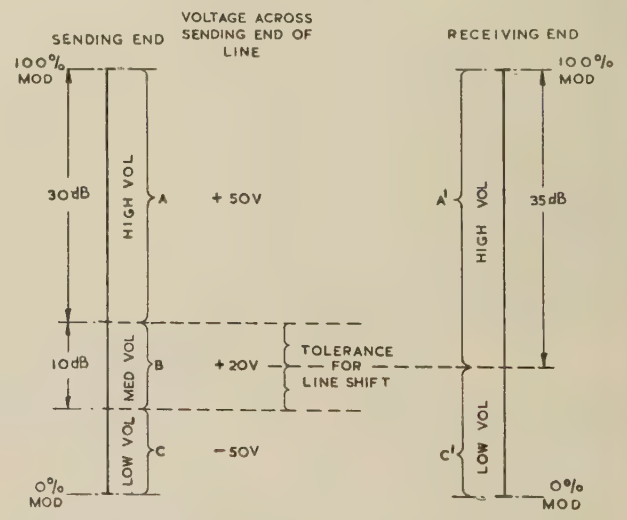


Fig. 1.—Level of programme signal at which detectors operate.
This differs for high and low frequencies, since aural grading is introduced in the detector circuits.

over the earth-return phantom circuit which has been derived from the programme circuit. Fig. 2 shows the general arrangement.

At the distant transmitter the programme signal is processed in a similar way to indicate high and low level. These level signals, derived from both the sending and receiving ends of the system, are compared, and so long as they coincide in nature the line and associated amplifiers are considered satisfactory. If, however, the d.c. signals at any time indicate an appreciable difference in programme level at the two points, the line executive monitor prepares to shut down the transmitter.

It should be mentioned that, although three signals are derived at the sending end, only the high and low level signals are selected at the remote transmitter. The medium level signal acts as a guard against small changes in transmission equivalent on the system, which could cause the one end to register high while the other was indicating low: Fig. 1 makes this clear.

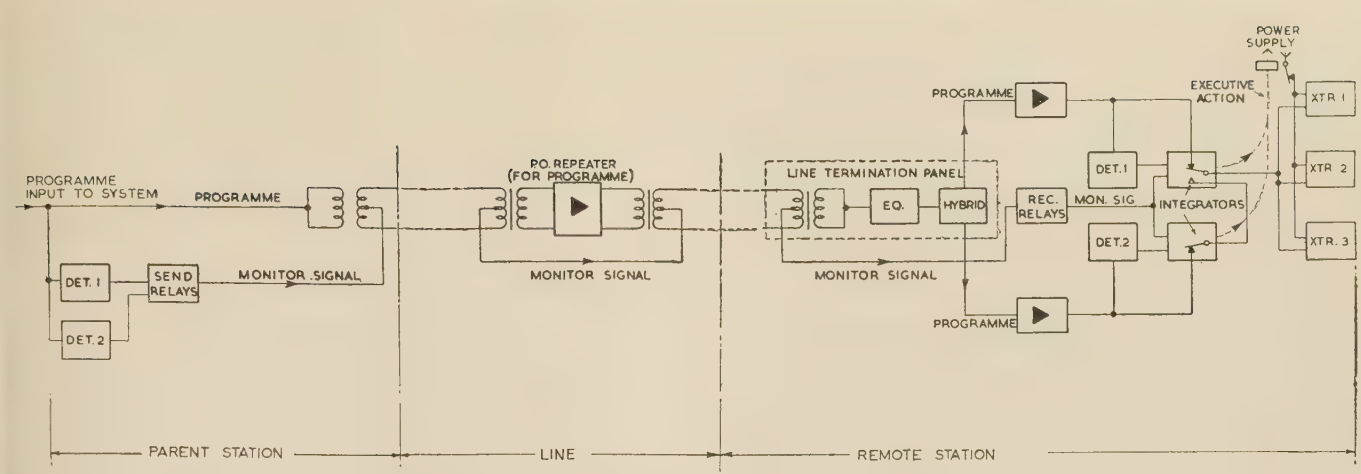


Fig. 2.—General arrangement of line executive monitor.

(4.2.1) **The Debit–Credit Integrator.**

An unbalance or lack of coincidence (debit) between the signals causes a capacitor to charge. The basic circuit for this integrator is shown in Fig. 3. It will be seen that by a multiple integration arrangement a number of short-duration faults or debits, in quick succession, will have only a slightly greater effect than one such fault. On the other hand, the same number of debits spread over a longer period will bring the integrator near or up to the close-down condition. The integrator is provided with a leak so that the build-up to a predetermined value takes place only if the fault condition is relatively persistent. If this predetermined value of integration is attained, the transmitter is shut down for a suitable period.

The integrator then automatically reverses in function, so that it has to be charged on the balance, or credit, condition to restore the transmitter to service. This will take place in a predetermined time, provided that no further faults—debits—arise. This provides a certain discretion in the executive action of the monitor. It is in a sense a 'reputation machine', not unlike a bank manager who becomes cautious in providing funds for the customer who has recently had an overdraft.

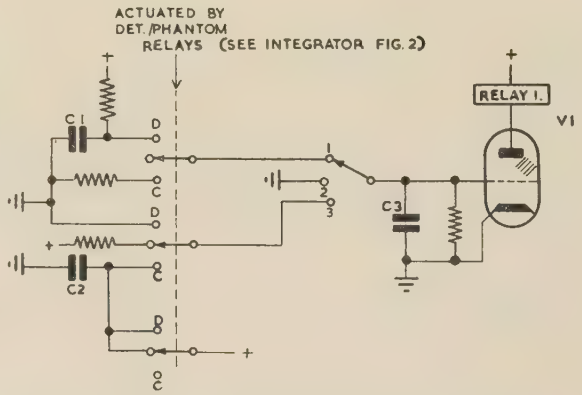


Fig. 3.—Simplified circuit of debit-credit integrators.

First strike of V_1 shuts down transmitters and switches circuit from 1 through 2 to 3.
Next strike of V_1 starts transmitters and switches circuit from 3 through 2 to 1.
When 2 is passed, C_3 is discharged and V_1 stops conducting.
C and D denote connections for credit and debit actions respectively.

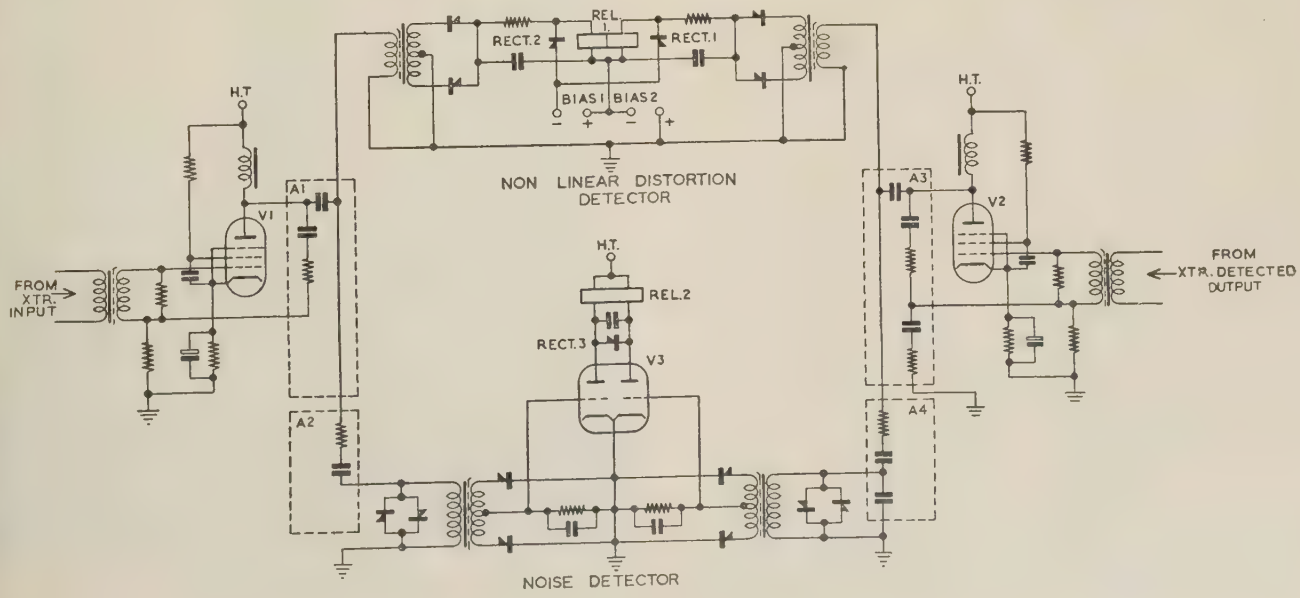


Fig. 4.—Circuit of transmitter executive monitor.

It is appreciated that it would not be sensible to close down a transmitter of this kind on a succession of closely spaced short line breaks. Only when the trouble has reached fairly serious proportions should this occur. The period of shut-down should then be such that restoration takes place only after there is a reasonable chance of freedom from further trouble, since repeated short interruptions to programme transmission can be very annoying to listeners. This is particularly so when it is remembered that in a large number of cases the same programme can be received, but not so well, from alternative transmitters.

(4.2.2) **Security of Executive Apparatus.**

Precautions have also been taken to ensure that a fault in the automatic apparatus is not likely to close down the station. For example, the detectors which derive the level information from the programme signals are duplicated. If they do not stay in register, i.e. both provide high or low levels at the same time, they are automatically precluded from taking executive action and an alarm is given at the nearest continually-staffed centre.

(4.3) The Transmitter Executive Monitor

The transmitter units at remote stations are watched by a simple type of monitor, supplemented by the usual protection arrangements. This type of monitor is relatively sensitive to faults peculiar to transmitters, i.e. hum or overload distortion. Overload distortion is assessed by comparing the input amplitude with the output amplitude when the modulation depth approaches 100%. The monitor is biased to be insensitive to amplitude distortion beyond this depth of modulation, since occasional peaks in programme signals will inevitably cause compression on the peaks of modulation. The general schematic of the monitor is shown in Fig. 4.

One monitor is provided for each transmitter unit, separation at the combined output being attained by special aerial combining networks.*

(4.4) The Automatic Telephone Indicator

In the event of the transmitter being completely shut down, either by the action of these executive monitors or by its protection circuit, an alarm is sent to the manned controlling station. For this purpose the monitoring channel is used in the reverse direction for the period of the alarm signal. When this alarm signal is given, technical staff at the controlling station may deduce the cause of the shut-down by making a telephone call to the distant station over the normal Post Office public telephone system. The distances involved mean that such a telephone call is usually handled by trunk or toll operators. The Post Office have co-operated in making provision for the fact that there is normally no one at the called station to answer the local exchange operator. Instead, when the operator establishes the required connection and the normal ringing tone has been heard for a few seconds, the station makes its automatic reply by a series of coded 'pips' which indicate the state of the station and the likely cause of the trouble. The 'pip' may be of long or short duration and high or low in pitch; this provides an adequate number of permutations, as shown in Fig. 5.

The same telephone facility can be used by the mobile maintenance teams whilst attending other distant transmitters. Should anyone be in attendance at the station when such a call is made, he can pick up the telephone and answer in the normal way.

(4.5) The Carrier Failure Alarm

As was forecast in 1949, most of these stations at which executive-type apparatus is employed have a part-time technical

* Morcom, W. J., and Bowers, D. F.: 'The Design of High- and Low-Power Medium-Frequency Broadcasting Transmitters for Automatic and Semi-Attended Operation', *Proceedings I.E.E.*, Paper No. 2362 R, April, 1957 (see p. 540).

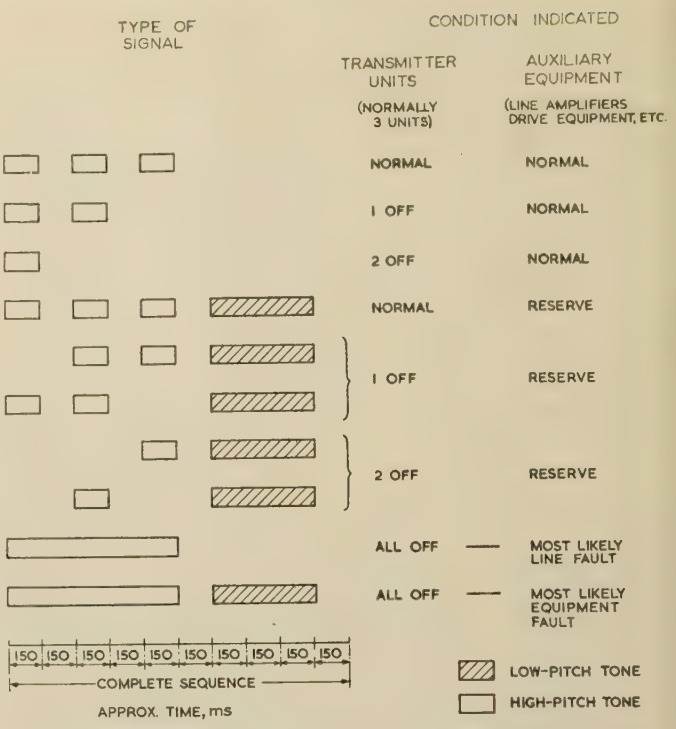


Fig. 5.—Signals given by the automatic telephone indicator.
The cycle shown is repeated for about 2 min.
Additional arrangements are available if required.

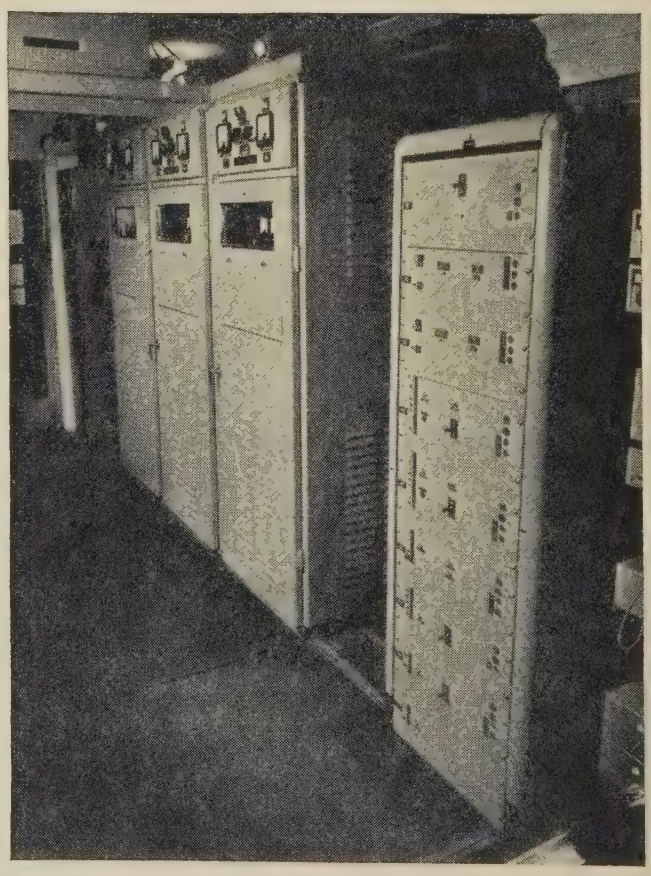


Fig. 6.—The 3-unit automatically controlled transmitter at Brighton.
The executive line monitor, including the debit-credit integrator, and the automatic telephone indicator unit are in the foreground.

assistant resident within the service area. A receiver in his home gives alarm if there is significant reduction in the strength of the local transmission. Since, in such circumstances, the receiver does not need automatic gain control, this circuit has been adapted to operate the relay which gives the alarm. If several transmitter units are normally radiating in parallel, the absence of one of them may be detected. A failure of the receiver itself produces an alarm, so that indication is always given of a fault at either the transmitter or the receiver. Outside transmission hours the receiver is switched off by a time switch.

Since the automatic monitoring equipment will close down the transmitter if any appreciable fault persists in the system, the resident technician is made aware of any serious trouble. For example, a break of sufficient duration in the line transmission will close down the transmitter, and the receiver alarm will operate. The alarm is of the locking type, so that it persists until it is manually reset.

By such an arrangement the technical assistant can lead a normal life without having to be always within earshot of his receiver. If he leaves his home he is required to keep in touch with the transmitter operation by telephone from time to time so that if the transmission has failed, he can proceed immediately to the station. Equipment at Brighton, a typical station, is shown in Fig. 6.

(4.6) The Extended Minor

This is really a form of the minor but made suitable for use on the earth-phantom monitoring channels. Like the minor it provides an alarm and is not normally used for executive action. It is used at the new v.h.f.-f.m. stations, on lines of up to 130 miles in length. Fig. 7 shows the general circuit arrangements.

In favour of paralleled transmitters the following principal arguments can be advanced:

(*a*) There is no break in the continuity of programme if one unit fails: in fact, reliability is similar to that provided by the twin or triple propellers of a ship. The reduction in power is within the range of the automatic gain controls now common in domestic receivers.

(*b*) The complete plant is continually giving service to the public, who should normally benefit from the whole rather than from the half.

In favour of two separate transmitters it can be claimed:

(*a*) The design of the station can be simplified, since no combining unit is required; this is replaced by simple impedance-matching networks and aerial change-over switches.

(*b*) Transmitters originally designed for attended stations can be used. Assuming that there is a complete standby transmitter, only a short break need occur if the operating transmitter fails.

Although opinions still differ on this subject, the B.B.C. decided to adopt the parallel-transmitter arrangement using output combining networks, so that a fault on one transmitter would not interfere with the working of the others. This also provided the required degree of separation so that the transmitter units could be monitored independently. A choice had then to be made of the number of units to be paralleled. Most of the semi-attended stations are of the relay or booster type, installed near isolated towns or pockets of population where satisfactory reception of the main high-power transmitters is not possible for reasons of distance or interference. As may be seen from Tables 1 and 2, most of these relay stations have an output power within the 1–2 kW range, giving a service area of up to 10 miles' radius. For these low-power transmitters it was decided that three paralleled units would be the optimum compromise between

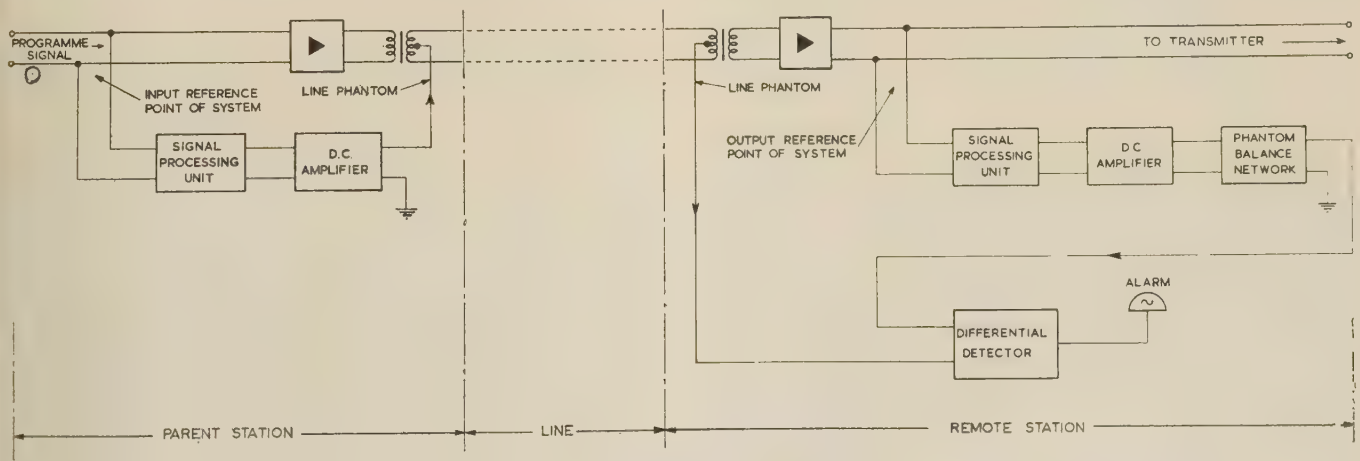


Fig. 7.—General arrangement of extended minor.

(5) TRANSMITTERS FOR SEMI-ATTENDED OPERATION

In the Address previously mentioned,[1] reference was made to a multi-unit experimental 1 kW m.f. broadcasting transmitter made up of ten 100-watt paralleled units. It was thought that, if these transmitters contained a comparatively large number of paralleled units, the chances of all of them failing simultaneously or in quick succession would be greatly reduced. In practice, these low-power units were found to be too small to permit orthodox transmitter design methods. The circuit complexity due to the multiplicity of units and the poor power conversion efficiency led to this design being abandoned. B.B.C. experience of this prototype multi-unit transmitter made it clear that the choice of transmitter arrangement therefore rested between two or three units in parallel, or one larger unit with another as standby.

reliability and complication. The design of a transmitter which has been used widely by the B.B.C. is described in the accompanying paper.*

So far, for most of the paralleled semi-attended m.f. transmitters, a power greater than 2 kW has not been required. These 3-unit transmitters have proved to be reliable in service, although they are rather more complicated and expensive than a single 2 kW transmitter. The predominant requirement of great reliability and safe rating for every component, including valves, has been carried into the design of these sets.

For low-power semi-attended transmitter operation (5 kW or less) it has been suggested that the susceptibility to breakdown, small though it is, would be further reduced if such transmitters were kept running continuously, i.e. not switched off at night or

* See page 540.

at times outside the scheduled programme hours. It has been claimed that weaknesses in components are more likely to be found under the various stresses caused by the application of voltage or changes in temperatures. So far, there is no evidence that such faults are increased in number by switching off the transmitters, and so there seems no justification for continuous running. These aspects are being watched very closely, and, of course, the procedure may be modified if the hours of programme transmission are extended.

The decision to adopt paralleled transmitters presented two alternative methods of automatic monitoring: either one monitor could examine each transmitter in turn and close down any which became faulty; or each transmitter could have its own monitor and be removed from service if its monitor disclosed a fault. In the former case it would be desirable for the monitor to be self-testing—there being only the one—before it rejected a transmitter unit from service. In the latter method, provided that the monitor is very robust, self-testing is unnecessary and the automatic circuits are correspondingly simpler. The simple robust transmitter executive monitor described earlier was therefore designed for this purpose.

The output combining circuits, essential for separation against fault conditions, provide the means of monitoring one transmitter distinctly from another. Combining circuits have been designed so that full efficiency of power transference to the aerial could be automatically maintained with three, two or only one unit in operation.[4] However, since the transmitter units are extremely reliable, it has not been considered worth while to bring some of the more elaborate arrangements into operation, particularly since the combining network is virtually a bottleneck through which all power passes, and a fault in it produces a complete shut-down of the transmitter. The loss of transmission time due to faults on the combining units is usually less than 0·001%, and it is hoped that the use of improved components will reduce this figure even further.

(6) DAVENTRY REMOTE-CONTROL THIRD-PROGRAMME TRANSMITTER

As will be seen from the accompanying list of stations, a most important example of the application of both remote control and automatic monitoring is the 150 kW Third Programme transmitter at Daventry,[5] which provides the Third Programme service for 27 million people and it is therefore a key station. For this reason, and also because a staff of engineers are engaged on other work on the same site, special monitoring arrangements were devised. This transmitter is in the building which used to accommodate the 25 kW 5XX long-wave transmitter and has been the principal radiator of the Third Programme on a frequency of 647 kc/s since April, 1951. Its performance during the last 5½ years is typified by its breakdown record from the 1st September, 1955, to the 31st August, 1956: during this period the transmitter was on half power (i.e. only one unit radiating) for 0·166% of transmission time, but only for 0·005% of the transmission time were both units out of action. These figures refer to transmitter faults and do not include interruptions due to power-supply failures, or faults on external programme lines.

One maintenance engineer only has been added to the staff of the Daventry short-wave station, and his duty is to perform protective maintenance five days a week when the Third Programme transmitter is not in operation. Monitoring is carried out entirely by automatic monitors, one being attached to each of the two paralleled transmitters. If either transmitter becomes faulty, it will be closed down by its monitor after the monitor has automatically tested itself and proved that it has detected the fault correctly. When one-half of the transmitter is closed down, the monitor on the other half is automatically denied executive action. It can merely report observed trouble by alarms to the nearby attended building. This arrangement gives the staff an opportunity of judging whether, despite faulty transmission, it is not better to maintain this key transmitter in service.

Again, the aerial combining unit has been designed so that, although the transmitters supply power in parallel, the separation between the outputs is sufficient for high-grade monitoring to be maintained over the individual transmitters. Another automatic monitor is fitted to cover the overall system which includes the transmitters, the feeder line and mast radiator, which is 1¼ miles away.

Recycling overload trips are fitted, and under power-fault conditions remove the h.t. supply and re-apply it a number of times before finally closing down the transmitter concerned. This prevents the transmitter being closed down and remaining closed down after a short-duration overload.

Although the aerial combining network is such that the one transmitter cannot affect the other when only one transmitter is working, the total radiated power is at first only a quarter of the normal full power, because half the transmitter power is dissipated in the combining network resistor. After 15 sec, however, if the one transmitter remains out of service, r.f. contactors select the working transmitter and connect it direct to the aerial feeder line, so that no power is wasted.

(7) ANALYSIS OF FAILURES

The causes of total shut-down of the m.f. semi-attended transmitters for the year 1st September, 1955, to 31st August, 1956, are classified into three groups, namely

(*a*) Line faults: this refers to faults on the final line section feeding the station from the main network; it does not refer to other lines earlier in the main network.
(*b*) Failure of the public electricity supply to the station.
(*c*) Local faults, i.e. those within the transmitting station, including the aerial system, etc.

(7.1) Line Faults

Line faults include breaks in transmission, noise or crosstalk and variations in response. Obviously, the short unloaded unrepeatered pairs such as those used to connect a transmitter to staffed premises a few miles away, are usually very reliable. Unfortunately, more trouble is experienced with the longer circuits, particularly if they are repeatered. In fact, records indicate that the fault liability on a given line is proportional, not only to its length, but is also (approximately) to the number of access points. The transmission time which is lost, or believed to be lost, owing to line faults is 0·052%.

Line faults are one of the main causes of shut-down, and, unfortunately, the nature of the faults tends to aggravate this trouble. Short but frequently repeated breaks are more annoying to listeners than a single long break. It is felt that this fact should influence the design of programme circuits and the precautions which should be taken in their routine maintenance. Since the automatic monitors are designed to be sensitive to all the various kinds of transmission interruption or distortion that might impair the programme signal, there is no simple way of distinguishing the kind of fault that has initiated an operation of the monitor unless an observer is available. It therefore follows that, at those semi-attended stations at which the monitor takes executive action, there is no simple way of determining the exact nature of the fault that caused a shut-down. However, since the monitor for the line and its associated apparatus is often separate from the transmitter monitors, it is possible to determine the broad reasons for the shut-down, e.g. whether the fault is on the line or transmitter, etc.

In at least one country on the Continent, radio broadcasting

receivers are considered to be more reliable than line feeds, and programme is distributed to the chain of stations by each obtaining its programme by radio reception from the previous station. Line feeds are installed, but primarily as a standby to the radio rebroadcasting links.

In the United Kingdom up to now, lines have provided the normal means of feeding the programme signal to the transmitters, and radio receivers are installed only as a standby. This radio-link reserve arrangement is easily applied to the Light and Third Programmes, which are never detached from their sources for local-network programmes. On the other hand, at different periods of the day some of the Home Programme transmitters are detached from the network and broadcast local programme material. This to some extent precludes the use of automatic switching to radio receivers for reserve purposes on the Home chain, for the wrong programme might be received unless special supervisory arrangements were made. When the v.h.f. network is fully established, the B.B.C. intends to use radio rebroadcasting more widely.

(7.2) Power Failures

The loss of transmission time due to power failures is 0·063%. The electricity authorities have co-operated very closely with the B.B.C. in maintaining supplies to these stations. They notify the resident technician when it is intended to carry out work at substations, or on power lines, which will interrupt the transmitting station supply. The figure given above is unusually high, for during the year under review there were several very long breaks in the power supply to the stations. From experience so far, the amount of breakdown due to power failures is insufficient to warrant the cost and complication of providing automatically-started standby engine-generator sets at these small stations.

(7.3) Local Faults within the Station

Wire aerials, usually suspended between two masts, are normally used. There has been little trouble from icing, storms or lightning, and the cost of erecting mast radiators is not warranted on a reliability count in the United Kingdom. The aerial, its masts and feeders are, of course, components in the transmission chain which would cause a complete shut-down were they to fail, but it is not practicable to duplicate the aerial system. For example, to provide a complete standby aerial on separate masts, a larger site at increased cost would be required and it would be difficult to provide the required polar diagram for both aerials. Failures due to masts, aerials and feeders represent a loss in transmission of only 0·001%.

Faults in the combining units account for a transmission-time loss of 0·000 7%.

Long-life valves are used, and valve failures account for a loss in transmission time of only 0·000 1%.

Faults due to transmitter components including auxiliary switchgear, etc., but excluding the combining units, are responsible for a loss in transmission time of 0·002%.

Faults in programme input and automatic monitor equipment, shutting down the station when otherwise it would have radiated, account for a time loss of 0·003%.

Human errors, including failure of the nearest manned centre to switch a remote-controlled transmitter and failure to set up clocks when there are changes in programme schedules or in changes to and from G.M.T. and B.S.T., account for a time loss of 0·002 5%.

A loss in transmission time amounting to 0·007 7% is due to faults occurring at an earlier stage in the system and also to unknown causes. From the above figures it will be seen that, of the total breakdown in transmission of 0·132%, only 0·009 3% is due to causes local to the transmitter. The relative proportions of the various causes of breakdown are shown in Fig. 8.

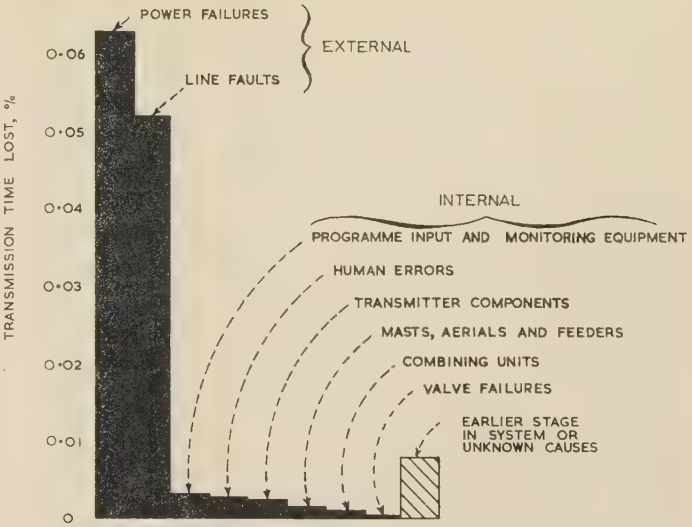


Fig. 8.—Total breakdown time from 1st September, 1955, to 31st August, 1956.

(8) TRANSMITTER MAINTENANCE

All transmitting stations require two kinds of attention, namely first-aid repairs and protective maintenance. The former are required to restore transmission after a breakdown and are usually performed by the resident technician after the automatic alarm has warned him that something is wrong. This maintenance is of a very simple kind, consisting of changing valves or components showing signs of becoming faulty. For safety reasons the extent of the first-aid repairs carried out by the resident technician, working alone, is limited.

Protective maintenance, including the measurement of transmitter performance, is carried out at attended stations by engineers who are trained for that function. At the m.f. semi-attended stations this requirement is met by the provision of mobile maintenance teams, of which there are three—one based at Bartley, near Southampton, responsible for all the semi-attended stations in the South of England and South Wales, one based at Stagshaw, near Newcastle upon Tyne, responsible for all the stations in Scotland and N.E. England, and one based at Moorside Edge, near Huddersfield, responsible for the stations in N.W. England and North and Central Wales. Each team consists of two engineers. At their base station equipment already exists to permit components or spares to be examined thoroughly and power-tested if necessary. Each team is provided with a self-drive vehicle which contains comprehensive testing equipment for full transmitter-performance measurements; a selection of appropriate spare units and components is also carried.

Eight stations are maintained by each of these three teams. The necessary maintenance work, together with travelling time, allows each station to be visited every four to six weeks. The resident technician is called in with the maintenance team when they are at his station, so that he can report to them anything suspicious and take careful note of any changes or readjustments which have been made. He therefore knows what parts of the equipment have been disturbed, and the station is handed over to him in proper working order before the maintenance team leave. If a fault develops which the technician cannot himself clear, he can contact the maintenance team through their base station. Often this results in assistance and advice being given which enables him to clear the fault condition.

In addition to providing specialist maintenance, the mobile

teams make regular m.f. field-strength measurements of semi-attended and attended stations at specific points during their journeys from one station to another. This is to ensure that no unsuspected fault has developed in the aerial system of a station and that the main high-power stations are maintaining their coverage. In this way it is possible to keep a regular watch on the field strength of B.B.C. m.f. transmitters at many points in the United Kingdom.

Before the mobile maintenance teams were initiated, m.f. semi-attended stations received both first-aid and protective maintenance from engineers at the nearest centre. A notable improvement in the breakdown records of these stations resulted from the use of maintenance engineers specializing in this work, compared with staff who are normally engaged on other engineering duties.

Since the majority of the semi-attended m.f. transmitters operate in synchronized groups, it is necessary to ensure that their carrier-frequency stability is maintained within $\pm 2 \cdot 5$ parts in 10^8. This is effected by using a crystal drive having this order of frequency stability in the presence of normal ambient-temperature and mains-voltage variations. The frequency of this drive is compared with a reference tone derived from the Droitwich 200 kc/s transmission, and manual adjustments are made by the resident technician if the frequency error displayed exceeds the requisite amount. It has been found that adjustments to the drive are necessary only at weekly intervals, except where there has been an abnormally large change in mains voltage or ambient temperature.

(9) RESIDENT TECHNICAL ASSISTANTS

At present, 20 m.f. stations are each supervised by a part-time technician resident in the service area. These resident 'watch dogs' are on full-time contracts, but are normally required to visit the station only for one hour a day, five days a week. Temporary reliefs are provided to cover sickness, leave, etc. It is gratifying that the staff engaged on these local duties are mainly pensioners or people who can only perform light work. B.B.C. staff on pension welcome the opportunity to do this work, for it provides an addition to their income, and they have been found excellent for this purpose. Many of them are senior staff who have passed the normal retirement age, or staff who have shown themselves to be highly reliable in the B.B.C. service and, unlike younger men, do not object to a job without prospects.

(10) THE ASSOCIATION OF BROADCASTING STAFF

The Association of Broadcasting Staff has been fully consulted during the development of this form of automation, and they have accepted it without objection on the grounds that it relieves staff for the performance of work of greater responsibility, where they can gain more experience and enhance their prospects of promotion. Furthermore, the Association appreciates that the technique permits a number of additional stations to be put into service at much less operating cost to the Corporation. It is an object of the Association to promote the efficiency of broadcasting.

(11) APPLICATION OF THE TECHNIQUE TO V.H.F. STATIONS

The planning of the new network of Home, Light and Third Programme f.m. transmitting stations has already been described,[6] so that it is unnecessary to give a complete description of the arrangements at these stations. However, it should be mentioned for completeness that there are two paralleled transmitters for each programme service (six in all). This particular paralleled arrangement is attained by feeding the programme signal into a drive modulator, the output of which supplies two separate r.f. amplifiers; the outputs of these are connected to separate halves of the aerial system. Since there are the three services, each half-aerial is fed from three output amplifiers which are separated from each other by suitable combining networks. This arrangement is shown in Fig. 9.

Since these stations are normally co-sited with television transmitters, advantage has been taken of the presence of television staff to provide the necessary maintenance. Only one extra engineer has been added to the existing television establishment, to deal with the additional protective maintenance work required by these six transmitters. In most cases, no extra staff is employed to monitor these stations. Relief from monitoring duties at the station is effected by one of the following methods:

(*a*) Automatic sequential aural monitoring at a nearby station if there is already provision to monitor the same programme aurally.

(*b*) Automatic monitor minors are used if no facility as in (*a*) is available, and the line arrangement, etc., is such that these particular monitors can be used.

(*c*) The extended monitor minor is used if the station feeding the transmitter with the programme signal is so distant that neither (*a*) nor (*b*) is feasible.

(*d*) The transmitter monitor (without executive action) is used if the transmitter is situated at an intermediate point in a line route which is already being monitored elsewhere.

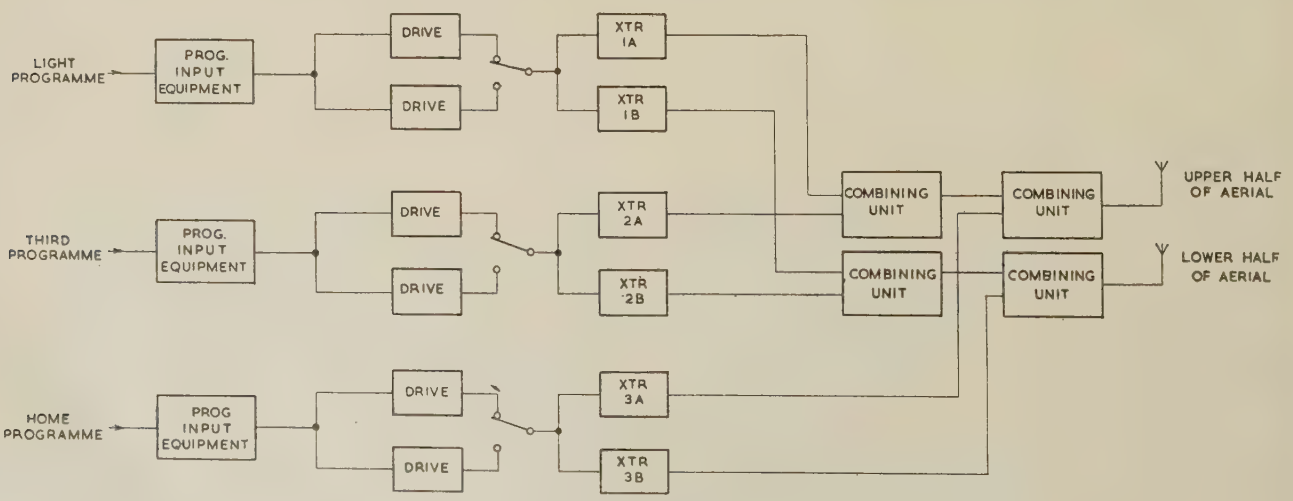


Fig. 9.—Transmitter combining arrangement.
A typical v.h.f. station with six transmitters.

If a fault is observed, a signal is given in the control room of the television building, where staff are continuously on duty during television programme hours. For a short period in the early part of the day, when the Home Programme is being radiated but no staff are needed for television purposes, special arrangements for alarm signals to be diverted to the nearest manned station are made. At these stations facilities are provided so that straightforward remedial changes may be made by remote control.

Since the television service requires only two shifts, the morning shift coming on duty at 8 a.m. and the evening shift going off duty at midnight, the Home Service f.m. transmitters must be started automatically at 6.15 a.m. each weekday and 7.30 a.m. on Sundays. This is done with time switches, and transmitter failure is observed at the remote monitoring station. The Light and Third Programme transmitters are started up and close down automatically in the same way, and local alarms are given if a fault occurs in any of the apparatus.

The single additional maintenance engineer can, of course, assist in the maintenance of the television equipment, and it may in turn be possible for him to get help from the television staff when it is desirable to have two men for maintenance of the v.h.f. transmitters. Staff are permitted to take out of service one of the two paralleled Home Service units, for routine maintenance purposes, before the evening-peak listening hours; this causes a 6 dB reduction in the effective radiated power. Nevertheless, every reasonable attempt is made to carry out major maintenance work during the period when there is no programme transmission. Routine maintenance of the Light and Third Programme transmitters is done outside their programme hours.

(12) APPLICATION OF TECHNIQUE TO THE OLD HIGH-POWER MEDIUM- AND LONG-WAVE STATIONS

In the period 1930–1945, the B.B.C. put into service a number of high-power m.f. transmitters, and one l.f. transmitter, for the radiation of the domestic services. Although still completely satisfactory in many respects, some of these stations are inefficient as power convertors, and by modern standards are generally large and cumbersome, having water-cooled valves and power supplies derived from rotating machinery. They are also expensive in staff. The modern semi-attended transmitters which could replace them are much more compact and could be fitted into the part of the building at present used for administrative offices, kitchens, etc., without interrupting the service. Since one man would normally be sufficient to supervise the operation of the transmitters, an establishment of only three engineers would be necessary, and they could be accommodated in houses built on the site. Maintenance could be performed by two engineers working at night or, if necessary, during the daytime when it is permissible to remove one half of the transmitter from service. Conversion of the stations in this way is profitable if the capital cost of making the change can be recovered in less than, say, 10 years, by the savings in staff salaries and power costs.

At the present time the margin in profit does not appear sufficient to warrant this being undertaken, particularly since the B.B.C.'s engineering effort is otherwise engaged. Nevertheless, this possible development is frequently reviewed.

(13) APPLICATION OF THE REMOTE AND AUTOMATIC CONTROL TECHNIQUE TO TELEVISION TRANSMITTING STATIONS

Although one or two lower-power and semi-attended television transmitters have recently been introduced on the Continent, none has yet been designed and put into service in this country. This is because the general design of television transmitters has not yet had time to settle down and the B.B.C. has wanted to give first priority to the building of the main high-power and medium-power television transmitters to extend the coverage of its service. Furthermore, the low-power transmitters have so far been at the ends of the network, e.g. Rosemarkie in Ross and Cromarty or Blaen Plwy in Cardigan, and have required close picture supervision because of the difficulty in maintaining good quality over the links which feed them. It has so far been necessary, therefore, for these small stations to reproduce the signal which they receive as a video signal and translate it back to a radio signal. The equipment of the station consequently becomes appreciably more elaborate and susceptible to breakdown. If the problem could be tackled by designing stations which simply receive the picture by radio at one frequency, convert it to another frequency and re-transmit it with whatever power amplification is necessary, the general design of unattended transmitters would be greatly simplified. It is premature to forecast how the technique will be extended to television stations, but it is obviously the next step and it is already being studied.

The practice of supplying the two halves of the aerial from two separate transmitter groups has been applied with success at the new London television transmitter at Crystal Palace.[7]

(14) DEVELOPMENT OF THE TECHNIQUE IN OTHER COUNTRIES

The European Broadcasting Union set up a working party to examine and, where possible, co-ordinate the development of this technique on an international basis. There are a number of semi-attended transmitters working on the Continent, and, as a result, information on new problems and development in each country is available and kept continuously up to date. The broadcasting authorities in Western Germany have made excellent progress in the application of the technique in the development of their extensive chain of f.m. transmitters.

(15) CONCLUSION

Although experience is still being gained and changes are being introduced, it is thought that the results so far obtained are generally satisfactory. The increased risk of breakdown, which is the risk that engineers in a public service are most reluctant to take, had to be accepted. The technique has saved about 200 men, representing more than £100 000 per annum in salaries, while maintaining an acceptable record of continuity which, it is hoped, may be improved still further. The essential requirement is that every piece of apparatus which is designed for use in semi-attended stations must individually have the greatest possible reliability. In the limit, completely reliable transmitters would need no automatic monitoring equipment and would therefore be free from faults on such apparatus. This form of automation has made it possible to provide a service to small pockets of population where it would otherwise be economically impracticable and also to provide nation-wide v.h.f.–f.m. coverage with the minimum of staff.

(16) ACKNOWLEDGMENTS

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(18) APPENDIX: ANALYSIS OF STATIONS WITH REMOTE OR AUTOMATIC CONTROL

Table 1

SEMI-ATTENDED M.F. TRANSMITTING STATIONS REMOTELY CONTROLLED

Station	Transmitter power	Service	Programme		Control		Breakdowns between 1st September, 1955, and 31st August, 1956		
			From	Distance	From	Distance	Local faults	External faults	Total
	kW			miles		miles	%	%	%
Liverpool ..	1	Third ..	Manchester	41	Manchester	41	0·005	0·082	0·087
Stockton ..	0·25	Third ..	Newcastle ..	35	Newcastle ..	35	—	0·008	0·008
Preston ..	1	Third ..	Manchester	29	Manchester	29	0·001	0·260	0·261
Daventry ..	150	Third ..	Daventry ..	—	Daventry ..	—	0·005	0·005	0·010
Dundee ..	0·25	Third ..	Edinburgh	75	Redmoss ..	66	0·001	0·044	0·045
Wrexham ..	2	Welsh Home	Cardiff ..	148	Penmon ..	85	—	0·067	0·067
Bournemouth	0·25	Third ..	Bartley ..	35	Bartley ..	35	0·010	0·009	0·019
Fareham ..	1	Third ..	Bartley ..	22	Bartley ..	22	0·008	0·104	0·112
Exeter ..	0·25	Third ..	Bristol ..	84	Plymouth ..	56	0·016	0·017	0·033
Towyn ..	5	Welsh Home	Cardiff ..	164	Penmon ..	79	0·008	0·120	0·128
Leeds ..	1	Third ..	Leeds ..	6	Leeds ..	6	0·068	0·008	0·076
Glasgow ..	2	Third ..	Glasgow ..	4	Glasgow ..	4	—	0·046	0·046
Edinburgh ..	2	Third ..	Edinburgh	5	Edinburgh	5	—	0·013	0·013
Newcastle ..	2	Third ..	Newcastle ..	5	Newcastle ..	5	} —	0·079	0·079
Newcastle ..	2	Light ..	Newcastle ..	5	Newcastle ..	5	}		

Table 2

SEMI-ATTENDED M.F. TRANSMITTING STATIONS AUTOMATICALLY CONTROLLED

Station	Transmitter power	Service	Programme		Breakdowns between 1st September, 1955, and 31st August, 1956		
			From	Distance	Local faults	External faults	Total
	kW			miles	%	%	%
Ramsgate ..	2	Basic Home ..	London	56	0·017	0·022	0·039
Barrow ..	2	Northern Home ..	Manchester ..	90	0·006	0·076	0·082
Cromer* ..	2	Northern Home ..	Manchester ..	250	0·002	0·210	0·212
Whitehaven ..	2	Northern Home ..	Manchester ..	150	0·028	0·120	0·148
Bexhill ..	2	W. of E. Home ..	Bartley	120	0·031	0·078	0·109
Brighton ..	2	W. of E. Home ..	Bartley	73	—	0·028	0·028
Folkestone ..	1	W. of E. Home ..	Bartley	172	0·005	0·321	0·326
Barnstaple ..	2	W. of E. Home ..	Bristol	120	0·005	0·287	0·292
Redruth† ..	2	W. of E. Home ..	Bristol	232	0·001	0·009	0·010
Dumfries ..	2	Scottish Home ..	Glasgow ..	90	0·101‡	0·279	0·380
Scarborough ..	2	N. Ireland Home	Stagshaw ..	112	0·010	0·185	0·195
Londonderry ..	0·25	Light Programme	Belfast	90	} 0·002	0·172	0·174
Londonderry ..	0·25	N. Ireland Home	Belfast	90			
Swansea ..	1	Third Programme	Cardiff	51	0·031	0·093	0·124

* In all cases alarm signals sent to station supplying programme with the exception of Cromer. Alarms for Cromer go to Postwick (25 miles).
† Semi-attended between 0630 and 0800 hours only.
‡ This figure is high, owing to two abnormally long-duration operational errors.

Table 3

SEMI-ATTENDED V.H.F.–F.M. TRANSMITTING STATIONS AUTOMATICALLY CONTROLLED

These transmitters have not been in service long enough for complete breakdown statistics to be given.

Station	Transmitter power	E.R.P.	Service	Programme	
				From	Distance
	kW	kW			miles
Blaen Plwy ..	10	60	Welsh Home ..	Cardiff	130
	10	60	*Light	Penmon	110
	10	60	*Third	Birmingham ..	100 (approx.)
Divis	10	60	N. Ireland Home ..	Belfast	6
	10	60	Light	Belfast	6
	10	60	Third	Belfast	6
Holme Moss ..	20	120	Northern Home ..	Manchester ..	21
	20	120	Light	Manchester ..	21
	20	120	Third	Manchester ..	21
Meldrum	10	60	Scottish Home ..	Burghead	50
	10	60	Light	Redmoss	33
	10	60	Third	Redmoss	33
North Hessary Tor	10	60	W. of E. Home ..	Plymouth	25
	10	60	Light	Bristol	90
	10	60	Third	Plymouth	25
Norwich	20	120	Midland Home ..	Postwick	18
	20	120	Light	London	140
	20	120	Third	London	140
Pontop Pike ..	10	60	Northern Home ..	Manchester ..	141
	10	60	Light	Newcastle	14
	10	60	Third	Newcastle	14
Sutton Coldfield ..	20	120	Midland Home ..	Birmingham ..	11
	20	120	Light	Birmingham ..	11
	20	120	Third	Birmingham ..	11
Wenvoe	20	120	Welsh Home ..	Cardiff	6
	20	120	Light	Cardiff	6
	20	120	W. of E. Home ..	Cardiff	6
Wrotham	20	120	Basic Home ..	London	30
	20	120	Light	London	54
	20	120	Third	London	30

* These transmitters are expected to be in service early in 1957.

Table 4

AUTOMATICALLY MONITORED M.F. TRANSMITTERS

Burghead	Scottish Home
Belfast	Third Programme
Bartley	W. of E. Home
Start Point	W. of E. Home
Westerglen	Light
Redmoss	Light
Moorside Edge	Light
Plymouth	Light

[The discussion on the above paper will be found on page 549.]

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THE DESIGN OF HIGH- AND LOW-POWER MEDIUM-FREQUENCY BROADCASTING TRANSMITTERS FOR AUTOMATIC AND SEMI-ATTENDED OPERATION

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SUMMARY

The paper describes two medium-frequency broadcasting transmitters, rated at 660 watts and 100 kW respectively, and their associated combining circuits. Three of the smaller units are used in parallel to give a 2 kW station with fully automatic control, while two 100 kW sets make up the semi-attended station at Daventry; however, to conform with international limitations, the B.B.C. operate them at 75 kW each, giving 150 kW carrier power from the two units in parallel.

Similar transmitters, for both medium-wave and long-wave broadcasting, have been supplied for service outside the United Kingdom.

(1) INTRODUCTION

The design philosophy of transmitters for unattended operation is given in the companion paper,* and there is general agreement that such stations should consist of a number of individual transmitters operating in parallel and feeding an aerial via a combining circuit which prevents interaction between transmitters and enables each transmitter to have its own protection and quality-monitoring circuits and to be unaffected by faults on associated transmitters. The Third Programme is radiated from Daventry on a frequency of 647 kc/s at a power of 150 kW, using two transmitters in parallel, and reliability was considered to be enhanced by such features as air cooling, a.c. heating of filaments, and an arc rectifier of improved design.

Since these transmitters have a high power output, it was decided that the combining circuit would employ automatically controlled h.f. contactors, so that in the event of a fault in one transmitter the combining circuit could be by-passed and the full power supplied direct from the working transmitter to the aerial.

For the low-power transmitters it was considered that three units in parallel represented the optimum arrangement for unattended operation. No moving machinery was used, all cooling being by convection, and metal rectifiers were employed. To make these small transmitters suitable for unattended use, relays are not used in the combining circuit, because the combining circuit is common to all the transmitters and a failure might thus put the station completely out of operation. The combining circuit is therefore of particularly robust construction to enhance reliability. Transmitter failure thus gives a field strength in direct proportion to the number of transmitters still functioning.

(2) THE 100 kW M.F. BROADCASTING TRANSMITTER

(2.1) Technical Performance

The requirements of the 100 kW m.f. transmitters are as follow:

(*a*) An a.f. input of 6 dB above 1 mW in 600 ohms for 100% modulation: an attenuator permits adjustment to the level normally adopted by the B.B.C.

(*b*) Power to feeder to be 100 kW carrier (maximum), operating at 75 kW at Daventry to comply with internationally agreed maximum permitted power of 150 kW.

(*c*) Maximum modulation of 100%.

(*d*) Audio-frequency distortion not to exceed 2% from 30 to 10 000 c/s for all levels of modulation.

(*e*) Frequency response to be ±2 dB from 30 to 10 000 c/s and ±1 dB from 50 to 7 500 c/s.

(*f*) Carrier stability and parasitic radiation to comply with current international regulations.

(*g*) Overall efficiency to be as high as possible.

(2.2) Preliminary Considerations

Earlier designs of high-power r.f. transmitters used much ancilliary equipment which needed frequent supervision by staff and considerable maintenance, and demanded excavation for the installation of generators, pumps, cooling systems, ponds, etc. The aim of the new design was to produce a transmitter which would not only run without close supervision, but would be easy to install and simple to maintain, and would take advantage of modern valves and improved circuit techniques to give better performance and efficiency than had previously been achieved. The medium-frequency transmitter was planned as a basic type from which long-wave and short-wave versions were later produced. The design of the transmitter permitted alternative layouts of the r.f. stages, while leaving the modulator, h.v. rectifier and control circuit common to all designs.

A transmitter with class-B modulation was chosen because it should retain its original adjustment without continuous attention better than any other system of operation, while giving performance and efficiency at least as good as other designs.

With the large valves cooled by air at low pressure and a single large industrial-type blower, a remarkable degree of reliability of the cooling circuit could be achieved: over five years of service no breakdown due to cooling troubles has occurred.

The choice of rectifiers was given careful consideration, and it was finally decided to use six AR61 excitrons—a glass-enclosed mercury-pool grid-controlled half-wave valve offering reliability, initial cheapness and easy replacement, with grid-control facilities for starting and overload protection.

Air-cooled triodes suitable for the modulator and r.f. stages did not exist prior to this design, and a new series, first with pure, and later with thoriated, tungsten filaments was developed; the compact filament construction permits a.c. heating to be used on both stages.

A single feeder and aerial system was chosen, and it was considered that reliability could best be served by installing two identical transmitters complete with their starting, monitoring and overload protection circuits and safety interlocks, and feeding the output into a separately interlocked combining unit. To give proper matching and freedom from interaction between transmitter outputs such a network must incorporate a dissipative element, so that, when one transmitter fails, the output from the other is divided equally between the feeder and the dissipative load. This is obviously a disadvantage, but is the price which must be paid for isolation between the transmitter outputs. The

* WYNN, R. T. B., and PEACHEY, F. A.: 'The Remote and Automatic Control of Semi-Attended Broadcasting Transmitters', *Proceedings I.E.E.*, Paper No. 2329 R, April, 1957 (see p. 529).

Mr. Morcom and Mr. Bowers are with Marconi's Wireless Telegraph Co., Ltd.

failure of one transmitter means that the power to the aerial falls to a quarter, i.e. the field strength drops 6 dB. This is not a serious price for the continuity of programme, but it was considered desirable to incorporate in the combining unit a series of contactors so that the serviceable transmitter could automatically be connected direct to the aerial.

The alternative to transmitters in parallel was to provide two transmitters, each of 150 kW, one working and one standby, but this permits only 50% utilization of plant and there is a break in programme during the change-over period.

Radio frequency drive to the two transmitters must, of course, be from a common source. Tests showed that the phase stability of the transmitters was sufficiently good that no form of automatic phasing was needed. A phasing amplifier of unity gain in one transmitter input enabled initial phase set-up to be made. The B.B.C. has arranged automatic change-over from the working to the standby drive to safeguard drive faults.

(2.3) General Arrangement of Transmitter

The main h.v. transformer, the modulation transformer and the main air blower are outside the transmitter enclosure, and the transmitter itself is enclosed in a surrounding framework (see Fig. 1). Inside this enclosure the a.f. circuits, the main and

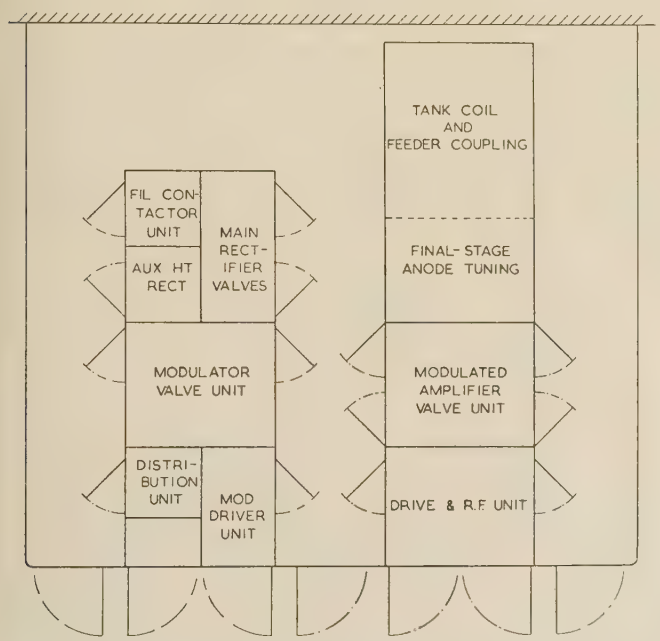


Fig. 1.—Plan of 100 kW transmitter enclosure.

auxiliary rectifiers, the control circuits and the contactors form one assembly on the left-hand side, with the r.f. stages making a second assembly on the right-hand side. Internal gangways give good access to all valves and components.

A fully sequenced pushbutton control panel with overload fault-indication lamps is mounted on the door to the central gangway. A duplicate panel mounted on a control desk gives extended control, but for remote control an auxiliary unit employing a few Post Office type relays is used.

At Daventry the two transmitters are arranged side by side with a space between in which the combining circuits are mounted.

(2.4) Cubicle and Enclosure Construction

The transmitter structure uses a novel system by which the whole assembly is built from a relatively small number of standardized parts. Aluminium-alloy panels formed in a box section can be bolted together to make a cubicle, or can be used with pillars and rails of complementary section to make larger cubicles or enclosures. This system is applied to a wide range of equipment, and it has made possible the use of standardized fittings such as hinges, handles, windows and interlocks. The cubicles are air-sealed by the use of rubber gaskets of special section.

(2.5) Interlocks

The equipment is mechanically interlocked so that personnel cannot come into contact with dangerous potentials. The locks on the door frame consist of a body carrying four plungers which, when free, engage in slots in a metal segment and hold it in that position. The door carries a plate with four pins which depress the plunger when the door is fully closed. The segment can then be moved into position to lock the door, and only then can the earthing switch be released and power applied. Access to the combining unit enclosure is by a door interlocked to the safety system of both transmitters.

(2.6) Circuit Design

(2.6.1) Modulation and A.F. Stages.

The modulator uses the high-level class-B system of modulation and is capable of producing excellent fidelity at all levels of modulation throughout the a.f. spectrum. The main modulator (Fig. 2) uses four air-cooled triodes of the same type as the final r.f. stage. Valves of type BR126, with pure tungsten filaments, and of the more recent type BR161, with thoriated-tungsten filaments, have been used, the latter raising the overall efficiency of the equipment by a reduction in filament power. A further operational economy became possible when the BR189 triode was introduced, since two of these valves operate very successfully in this position. Comparative data for these valves are given in Table 1.

Table 1

CHARACTERISTICS OF AIR-COOLED TRIODES

	BR161	BR189
Filament voltage, volts	9	9
Filament current, amp	175	240
Peak cathode current, amp	45	70
Peak anode voltage, kV	12	12
Peak anode dissipation, kW	15	27
Peak grid dissipation, kW	1	1·25
Mutual conductance, mA/volt		
at $V_a = 9$ kV	45	34
at $I_a = 1{\cdot}2$ amp	23	—
at $I_a = 2{\cdot}0$ amp	—	27·5
Inter-electrode capacitances, pF		
anode-grid	36	51
anode-cathode	1·5	1·5
grid-cathode	57	68

The output circuit of the modulator, comprising modulation transformer, modulation reactor and blocking condenser, is designed to operate with very little distortion over the range of 30–10 000 c/s. No real problems are encountered at mid-audio frequencies, where the modulation components perform in a classic manner. The requirements at low and high audio frequencies are conflicting: the high inductance of the modulation transformer normally needed to give good results at 30 c/s would restrict performance at high audio frequencies, owing to the unavoidable resonance of the leakage inductance and stray capacitances. In this transmitter an effective compromise between the conflicting requirements was made by treating the modulator output circuit as a π-filter. At low frequencies the shunt arms of the filter are formed by the secondary winding of

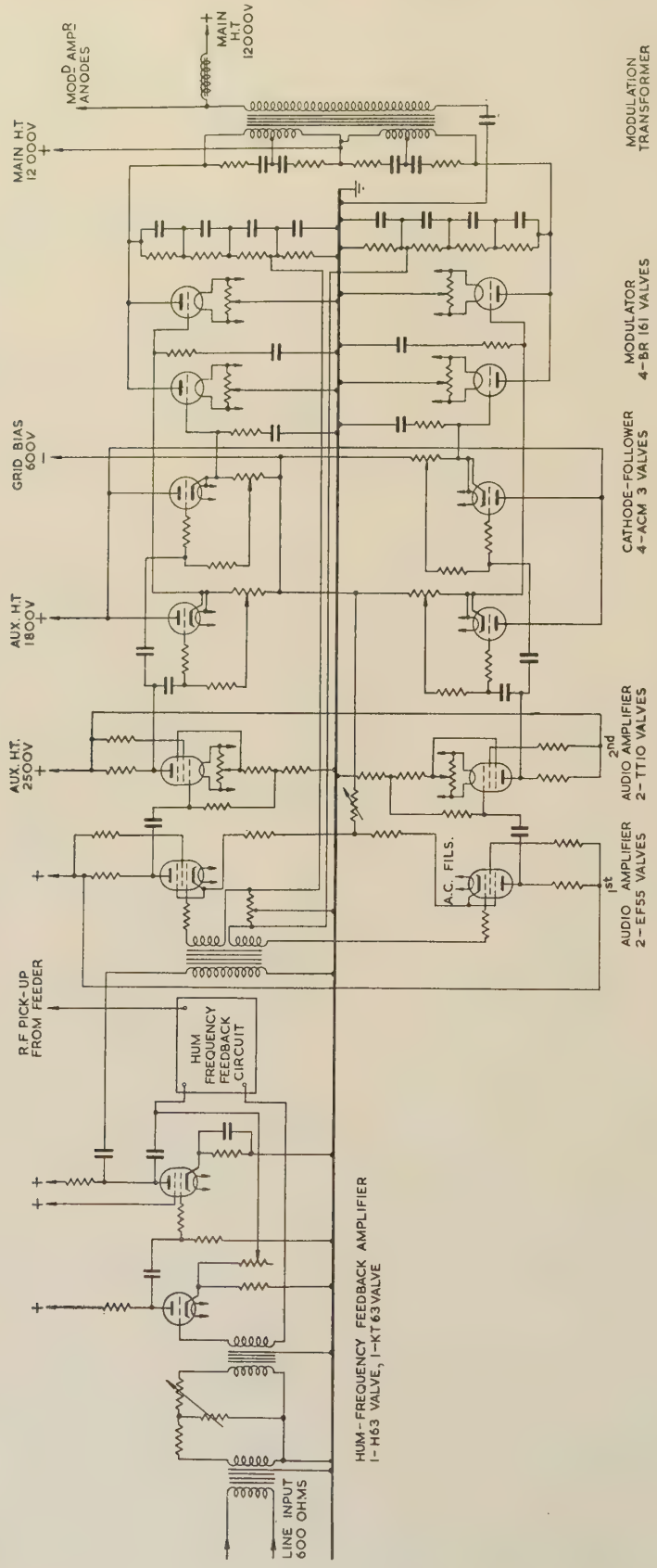


Fig. 2.—Modulator stage of 100 kW transmitter.

the modulation transformer and the modulation reactor and the series arms by the blocking condenser. At high frequencies the shunt arms comprise the capacitance to earth of the modulation transformer and the r.f. circuits, while the series arm is formed by the leakage inductance of the modulation transformer. As a result of this treatment it has been possible to present the modulator valves with a substantially non-reactive load over the full a.f. range. The distortion is maintained at a low figure over the full range of audio frequencies by the use of negative feedback, which is still fully effective and stable at 10 kc/s. This has been achieved by minimizing the phase shift throughout the modulator chain. A contributory factor is the use of *RC* damping units in shunt with each primary winding on the modulation transformer—an arrangement preferable to the more conventional low-pass filter in the modulator output circuit, which would cause a more rapid change of phase at the harmonics of the higher audio frequencies and hence limit the degree of feedback which could be applied without instability. The *RC* damping units effectively reduce the peaks in response at supersonic frequencies and thereby reduce the tendency to self-oscillation. The modulator operates with 17 dB of feedback and is normally tested with an additional 6 dB, but laboratory tests were made with up to 30 dB of feedback to check stability.

The main modulator is driven by a direct-coupled cathode-follower using four ACM3 triodes. With this circuit the grid current to the main modulators passes through the cathode-follower valves from the anode supply to the stage. Thus, with a supply of adequate regulation, the cathode-follower is the only cause of distortion of the grid driving voltage. The ACM3 has high mutual conductance, and as a cathode-follower has an output impedance of 30 ohms.

The ACM3 has an oxide-coated filament, was originally designed for pulse operation and must not be driven so that grid current flows. In consequence, the circuit embodies biased diodes to prevent positive grid drive, and the time-constants of the smoothing circuits of the associated power supplies have been arranged to prevent positive grid transients during switching of supplies. The diodes have a very sharp cut-off and can be preset to limit modulation.

The cathode-follower is preceded by two push-pull *RC*-coupled amplifiers designed on wide-band principles to facilitate the application of feedback, which is applied to the grids of the first push-pull amplifier valves from *RC* potentiometers connected between the anodes of the modulator valves and earth.

The programme input is applied to the transmitter at +6 dB level for 100% modulation, an attenuator permitting the gain to be adjusted to suit the normal B.B.C. requirement of +12 dB for 100% modulation. A 2-stage amplifier provides the gain necessary to drive the first push-pull voltage amplifier, and also operates to reduce hum by applying feedback at the offending frequencies. The feedback voltage is derived by balancing a sample of the audio output of the transmitter against the audio input to the first push-pull amplifier stage.

(2.6.2) **R.F. Stages.**

On peaks of modulation, potentials of 50 kV occur, but the size of the final r.f. valve unit is kept small by adequate corona protection.

Distortion of the programme can be caused both by the modulator and the modulated r.f. amplifier. This restricts the design of the latter in two ways, since it can produce distortion by its own alinearity or by presenting a highly reactive load which causes distortion in the modulator.

Alinearity of the r.f. amplifier is minimized by using grid-leak bias and adequate grid-driving power. The grid-driving voltage is derived from capacitive potentiometers in the output of the

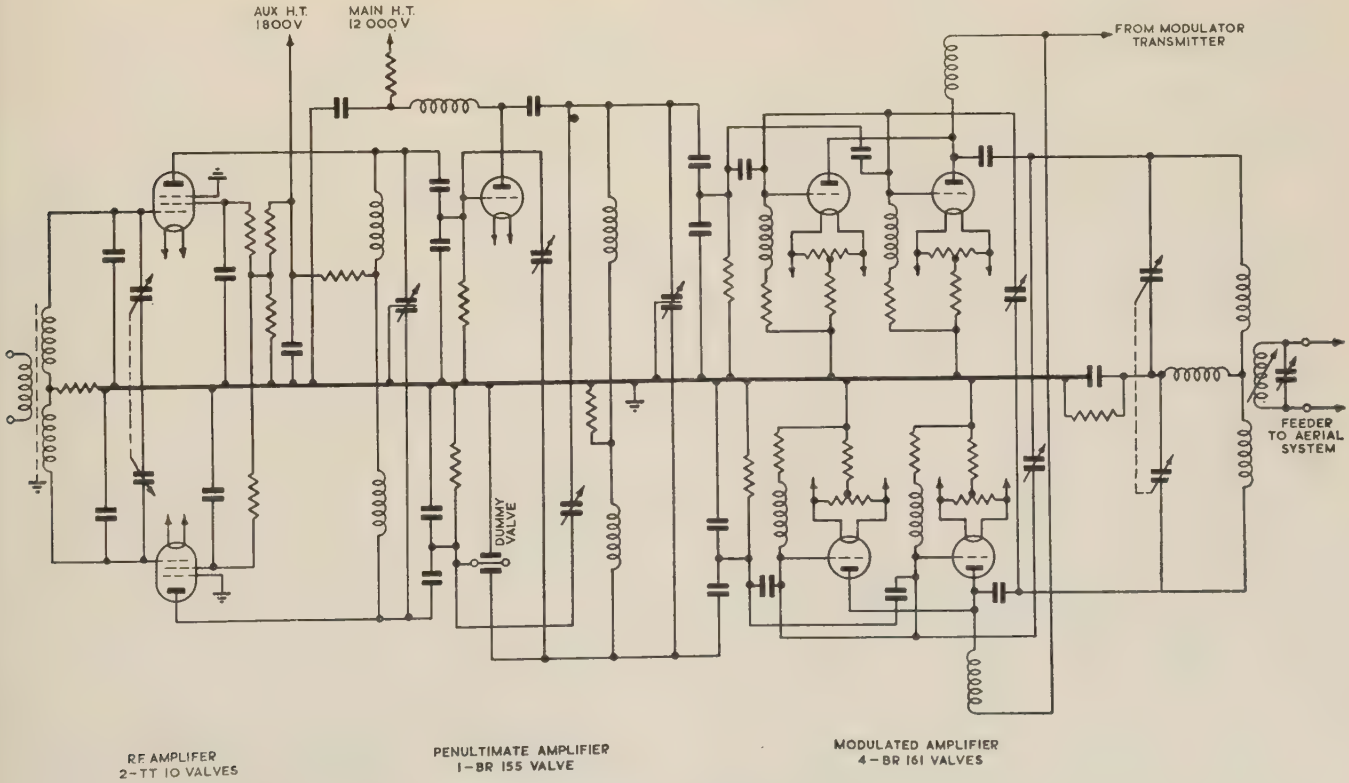


Fig. 3.—R.F. stages of 100 kW transmitter.
R.F. input is 5 watts at radiated frequency.

penultimate amplifier, which operates with loaded Q-factor of between 30 and 40 over the frequency band. This allows the grid-voltage waveform to be maintained up to 100% modulation, and also reduces slightly the hum due to the a.c. heating of valve filaments.

The reactive portion of the modulator load at high audio frequencies normally comprises the capacitance associated with the tank circuit together with valves, leads, etc. The capacitance of the tank circuit has been reduced by using a push-pull arrangement, as shown in Fig. 3, proportioned to present to the modulator output circuit the shunt capacitance needed to satisfy the requirements described in Section 2.6.1. The tank-circuit inductor and tuning capacitors are isolated from earth, the centre point of the capacitors being held at near earth potential by a relatively low-value capacitor, which provides a low-impedance path to earth for second-harmonic currents and caters for the out-of-balance currents on the fundamental frequency. It is shunted by a leak resistor which gives optimum voltage distribution across the various components in the tank circuit.

Medium-frequency broadcast transmitters generally use an unbalanced open-wire feeder of 200–300 ohms impedance for connection to the aerial. Where climatic conditions are adverse, concentric feeders of 70–100 ohms impedance give more reliable operation. This particular design of transmitter caters for the two conditions by using alternative versions of the feeder matching and coupling circuits, the Daventry transmitters employing the higher-impedance type. The feeder coupling circuits cover the range of nominal feeder impedance with adequate tolerance to cater for conditions such as those met when using phased aerial arrays, where it is often impossible to obtain the required radiation pattern without a degree of mismatch of the feeder at the transmitter.

It is of great practical advantage to be able to vary the power output of the transmitter easily, particularly if this can be done while radiating. This has been achieved by using variable magnetic coupling between the feeder and the final-stage tank circuits.

In the design stages several possible arrangements were investigated, the criterion being that which gives least capacitive coupling between tank and feeder coils for a given coefficient of coupling and voltage rating. The design used on this transmitter (Fig. 4) uses two coaxial tank coils with the feeder coil in the form of a flat spiral moving in and out between them. The final arrangement gives sufficiently low capacitance between coils to allow the output power to be varied over a range of 10 : 1 without retuning.

Variable capacitors are used to tune the final r.f. stage to ensure that the application of classic methods of tuning gives the requisite results. Two versions of this tuning unit have been used, one having air-spaced condensers and the other pressurized nitrogen-filled units.

The transmitters supplied to Daventry had provision for mounting four air-cooled valves in the final r.f. stage. The filament supplies and connectors are adequate for the use of four tungsten-filament valves (BR126), and later simple modifications enable four valves with thoriated-tungsten filaments (BR161) to be used in their place. Both types are satisfactory in operation and give very little difference in performance, apart from the saving of filament power. Later versions of this transmitter employ two BR189 triodes in this stage with further economy in running costs.

The push-pull earthed-cathode circuit is neutralized by a bridge employing fixed mica condensers.

The penultimate stage uses a single BR155 triode operating a push-pull circuit incorporating a dummy valve to maintain symmetry and provide a neutralizing adjustment.

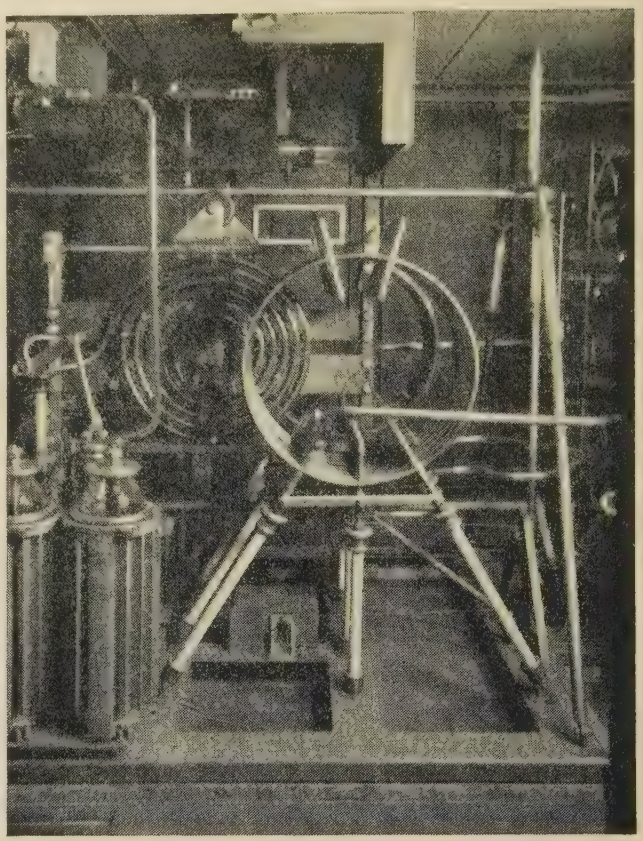


Fig. 4.—Final tuned circuit and feeder coupling circuit of 100 kW transmitter.

(2.6.3) **Main H.V. Rectifier.**

The 12 kV main h.v. rectifier is accommodated inside the transmitter enclosure, and consists of six glass-bulb mercury-pool grid-controlled half-wave AR61 valves arranged vertically in two rows of three; the cubicle is 5 ft wide, 2 ft 6 in deep and 6 ft high, each valve being on a sub-assembly, complete with its excitation and grid-control circuits. This type of rectifier has all the advantages of the conventional multi-anode mercury-arc valve, together with easier and cheaper replacement and servicing facilities.

Ignition of the rectifier is by a dipper electrode inside the tube, which is held in the working position by the leakage flux of an inductor connected in the ignition circuit and mounted close to the outside wall of the tube. The h.v. supply is applied in three equal steps by the sequential application of excitation voltages to the rectifiers, provision being made for stopping at any of these voltage intervals for tuning or testing. The h.v. supply can be reduced to approximately five-sixths of its normal value by shifting the phase of the supply to the grids, thereby delaying the firing of the main arc.

The overload control circuits of the transmitter incorporate a 3-shot restoration system operating through the one-third, two-thirds and full-voltage conditions: after three reclosures on a persistent fault the main circuit-breaker is tripped out.

(2.6.4) **Filament Supplies.**

The filaments of all the valves are heated by alternating current. The initial use of four type BR126 valves with tungsten filaments in the r.f. stage and four BR126 valves in the modulator determined the rating and the arrangement of the filament circuits.

The four valves in the r.f. stage are excited at 45° to each other

to reduce hum, while the main modulator pairs of valves on each side of the push-pull circuit are excited at 90° to each other to give cancellation of the hum components at the second harmonic of the a.c. supply. The filaments of the two pairs of valves are excited in phase, since the modulation transformer will then cause a large measure of cancellation of the resulting hum-frequency components in the anode currents of the valve. This arrangement gives a hum level of some −57 dB with reference to 100% modulation without any other correction being applied. These transmitters were later re-valved with thoriated-tungsten-filament triodes of type BR161, and still later editions use two BR189 valves in the r.f. stage and two in the modulator, resulting in a further improvement in efficiency and a reduction in valve cost. With these valves the r.f.-stage filaments are excited 90° out of phase and those of the modulators are in phase. This gives a hum level comparable with that obtained with the original four valves, owing to the much lower current requirements of the thoriated-tungsten filaments.

Inductors are used, for a short time duration, in each leg of the 3-phase supply to the filament circuits to limit starting current.

The a.c. supply for the filaments is obtained via an automatic voltage regulator which maintains an output voltage to a tolerance of ±1% for input variations of up to ±10%.

(2.6.5) **Overload Protection.**

In addition to the normal h.r.c. fuses or switchgear, overload protection has been used on the circuits as detailed below.

The high-power filaments have automatic-resetting overload devices in the 3-phase supply; they are fitted with oil dashpots and can be adjusted to operate between 125 and 200% of normal current. Operation of these devices causes all power other than that for the early stages to be removed from the transmitter.

The bias and auxiliary h.t. circuits use automatic resetting d.c. overload relays, operation of any one of which removes the associated a.c. power supplies and also opens the circuit-breaker of the main h.t. rectifier.

The main h.t. rectifier protection circuits operate at high speeds and use the arc-suppression properties of the AR61 excitrons to remove the anode voltage. Every major valve has an overload relay fitted in its cathode circuit, and a comparator relay monitors the input/output relationship of the final r.f. stage. On the secondary side of the h.t. transformer the main h.t. rectifier incorporates three overload relays of the high-speed telegraph type, with a transit time of less than 2 millisec; these apply arc-suppression to the six rectifiers simultaneously through a pair of thyratrons, and these in turn operate relays which cut off the supply to the primaries of the excitation transformers, thus extinguishing the excitation arc. At the same time an *RC* combination in the control-grid circuit of each rectifier applies a negative bias to the control grids, the time-constant of these elements being sufficient to ensure complete suppression of the main arc. The time from the commencement of a fault to the isolation of the anode voltage varies between 8 and 16 millisec and depends on the amplitude of the alternating voltage on the anode of the rectifier at the instant when arc-suppression is applied.

The comparator relay has two coils, one being excited from a shunt in the cathode circuit of the final-stage valves and the other from rectified output derived from a current transformer coupled to the outgoing feeder. The currents in the two coils are adjusted to be equal for normal working, and operation of the relay follows when this state is disturbed by a predetermined amount.

Protection of the main h.t. rectifier is afforded by three relays supplied by current transformers connected to the secondaries of the associated h.t. transformer. This enables the overload settings to be adjusted to fine limits without being disturbed by switching transients which occur in the primary side of the transformer.

The main h.t. rectifier includes circuits for 3-shot operation of the arc-suppression devices. A short-duration fault is cleared by removal of the anode voltage, and the programme is restored in 1¼ sec. For faults of a persistent nature the anode voltage is restored twice, and on the third fault indication the associated circuit-breaker is opened. Precautions have been taken to minimize the damage which might occur in the event of a flash arc (Rocky Point effect) on the main transmitter valves. Every iron-cored component associated with the d.c. supply system is fitted with surge diverters, connected in parallel with its windings and comprising spark-gaps and series-connected non-linear-resistance elements. Thus, if a flash arc occurs, the rate of rise of fault current is initially limited by the iron-cored component connected in series. This surge of current could produce a high voltage across the winding of the component, which would cause damage. The surge diverter is set to flash over at a safe voltage and the consequent arc is quenched by the non-linear resistor, which also serves to limit the current flow into the flash arc. In this particular transmitter the initial rise of current due to the flash arc operates the high-speed arc-suppression system to isolate the anode supply, leaving the energy stored in the condensers to be discharged. The surge diverters have been chosen to restrict the peak fault current to 200 amp.

(2.6.6) **Monitoring for Test Purposes.**

A visual and aural monitor is built into the transmitter; it uses a double-beam oscillograph for visual inspection of waveforms throughout the transmitter, pick-up points being provided on the output of the main h.t. rectifier, at points in the modulator and on all r.f. stages from the drive input to the feeder output. The sample voltages are selected and applied to the two beams of the oscillograph by two selector switches, the pick-up points being so arranged that comparisons of waveform can be made; alternatively the X-plates of the oscillograph can be switched to the a.f. input, in which condition an adjustable phasing circuit is used so that, with a trapezium pattern on the oscillograph, it is easy to appreciate a very low order of distortion.

Percentage modulation is indicated by a meter on the monitor, which also provides for an extension instrument for mounting on the control desk. This facilitates checking the transmitter during routine maintenance and forms no part of the overall automatic monitoring employed by the B.B.C.

(2.7) Power Consumption

The consumption figures for this transmitter are given in Tables 2 and 3.

H.T. supply regulation accounts for the slight fall of power on modulation.

Table 2

CONSUMPTION USING EIGHT BR161 VALVES

Condition	Total input	Power factor	Measured power in load
	kW		kW
Carrier	187	0·92	100
40% modulation ..	223	0·90	109·5
90% modulation ..	276	0·93	130
100% modulation ..	288	0·91	139

Table 3

CONSUMPTION USING FOUR BR189 VALVES

Condition	Total power input	Power output
	kW	kW
Carrier	174	100
30% modulation	195	104·5
40% modulation	203	108
100% modulation ..	257	142

(2.8) Modulation Performance

The assessment of the quality of audio devices has been the subject of numerous discussions, papers and controversies. It is, in fact, a delicate topic, particularly of recent years since the advent of the so-called 'high fidelity' audio reproducers has produced a class of highly critical listeners. Nevertheless, no piece of equipment can be fully assessed by listening tests, and some basis of measurement must be applied as a criterion of performance. At one time it was thought adequate to specify a.f. harmonic distortion as a measure of quality, but the inadequacies of this measurement became apparent and much work was done to discover a method of measuring audio performance which truly related to the listening tests. One method is to measure the intermodulation products, the procedure being to apply two tones of equal amplitude with a difference frequency of 1 kc/s. This test was applied to the present transmitters, and Fig. 5 shows the values obtained.

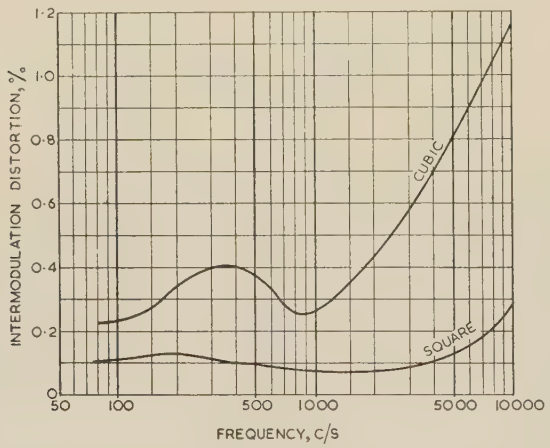


Fig. 5.—Intermodulation distortion in 100 kW transmitter: square and cubic terms.

Conditions: Modulation depth = 90%.
f_1–f_2 = 1 000 c/s.
Output power (carrier) = 100 kW.

In order to provide some empirical assessment the B.B.C. also passes transmitters through other subjective tests in order to ensure that the level of non-linear distortion is insignificant on normal programme transmission. It is interesting to note that with transmitters in parallel the resulting distortion is of the order of the mean value of the individual transmitters.

(2.9) Remote-Control Panel

The remote-control panel consists of a rack-mounted unit containing an on-off selector switch, an indicator lamp for each transmitter, a transmitter-failure warning bell and its muting switch and a combining-unit isolating switch.

The normal process is to switch the transmitter to 'on', and in due course as the transmitters are energized both indicator lamps light. Failure of one transmitter rings the alarm bell, the identity of the faulty transmitter being shown by the indicator lamps. At an appropriate moment the operator can press the switch which isolates the combining unit and connects the effective transmitter direct to the aerial. Alternatively, this operation can be performed automatically by preselection at the transmitters.

(2.10) Drive and Phasing Circuits

Two master-oscillator units, one working and one standby, are used, stability of 1 part in 10^6 being obtained by using 5° X-cut bar crystals in the band 66·7–100 kc/s. The tuned-anode maintaining valve operates at three times crystal frequency, and is followed by amplifiers and further frequency multiplication to give a 5-watt output at radiated frequency into a balanced 100-ohm load. Duplicate valves are used in all stages to minimize risk of failure, and change-over to spare drive is automatic.

The relative phasing of the two transmitters is by a single unity-gain amplifier in one transmitter input, and manual phasing was found to be adequate. Phase indication is given by measuring the current in the resistive arm of the bridged-T combining network.

This description of high-power transmitters for semi-attended working has dealt with details of the transmitter units. In the companion paper* a description has already been given of the auxiliary automatic monitoring and switching arrangements which the B.B.C. have added to the installation at Daventry to give further security by comparing the output signal of the installation with the signal at the input. The way this equipment interlocks with that of the transmitter, so that appropriate action is taken, is also described.

(3) 2 kW M.F. UNATTENDED BROADCASTING TRANSMITTER

(3.1) Technical Performance

The requirements for the 660-watt m.f. transmitters were as follow:

(*a*) An a.f. input of 1 mW in 600 ohms for 40% modulation.
(*b*) Power to be 660 watts per unit into 300 ohms (unbalanced).
(*c*) Maximum modulation to be 100%.
(*d*) A.F. distortion not to exceed 3% at 90% modulation at the limits of the a.f. range.
(*e*) Frequency response to be −2 dB between 50 and 7 500 c/s.

(3.2) Preliminary Considerations

A transmitter for semi-attended or unattended use must be able to continue to give service over long periods without attention. It is very evident that a station consisting of several identical low-power transmitters all contributing to the radiated power, each capable of its own individual monitoring and able to take itself off the air in the event of a fault without affecting the remainder of the transmitters or the quality of the transmission, would fulfil this requirement. The design of such a transmitter should have the accent on ruggedness and simplicity, and the following questions have an obvious bearing on the design problems:

(*a*) Can transmitters run in parallel over long periods without adjustment, and what, if any, special equipment does this entail?
(*b*) Can a faulty transmitter be detected and isolated without breaking the continuity of the programme?
(*c*) What are the main causes of failure of transmitters, and how can they be reduced?
(*d*) What is the economic number of transmitters?
(*e*) Where common-frequency working is necessary, how can constancy of frequency be assured?

* See page 529.

The result was the decision to design a 2 kW station comprising three identical self-contained 660-watt transmitters, with a common drive, a.f. input from a line amplifier and the output from each matched to feeder impedance. Each transmitter was to have its own overload protection and relay sequence starting circuits. Reliability was considered to be best served by using the minimum number of valves, with metal rectifiers for the d.c. supplies, and tests showed that the phase stability of a transmitter employing only two r.f. stages was remarkably good. To ensure drive stability for common-wave operation the B.B.C. provided a high-stability source of frequency which can be checked against a reference frequency.

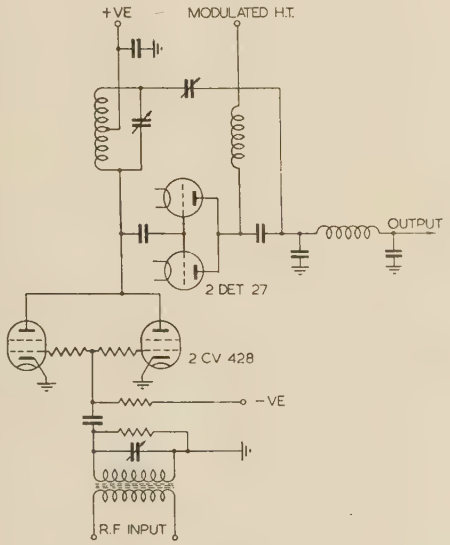


Fig. 6.—R.F. stages of 660-watt unit transmitter.

(3.3) General Arrangement of Transmitter

Each 660-watt transmitter—rectifiers, modulator, r.f. stages and control circuits—is contained in a cabinet, 7 ft high and 2 ft 6 in square, complete with safety interlock circuits. It is thus a self-contained unit, requiring only single-phase mains, a.f. and drive inputs, and will supply 660 watts to a feeder of 300 ohms unbalanced impedance.

The three cabinets are bolted together, and the r.f. outputs feed to a combining unit mounted on top of them. This arrangement has been used because of space limitation in the type of building in which these transmitters are housed. Any one of the three transmitter units may be serviced while one or both remaining units are in operation, and to assist servicing, only two types of valves are used in the transmitter.

The main r.f. and modulator valves are mounted on a deck at the top of the cabinet in such a manner that radiated heat can do no damage to the cabinet wiring or any components. This enables the transmitter to operate without mechanical cooling arrangements, which adds to the reliability. Below this level are mounted the low-power modulator stages, the B.B.C.-type monitor and the penultimate r.f. driving stage. Grouped in the bottom of the cabinet are the supply rectifiers, the h.t. transformers and smoothing systems, and the modulation iron-cored components.

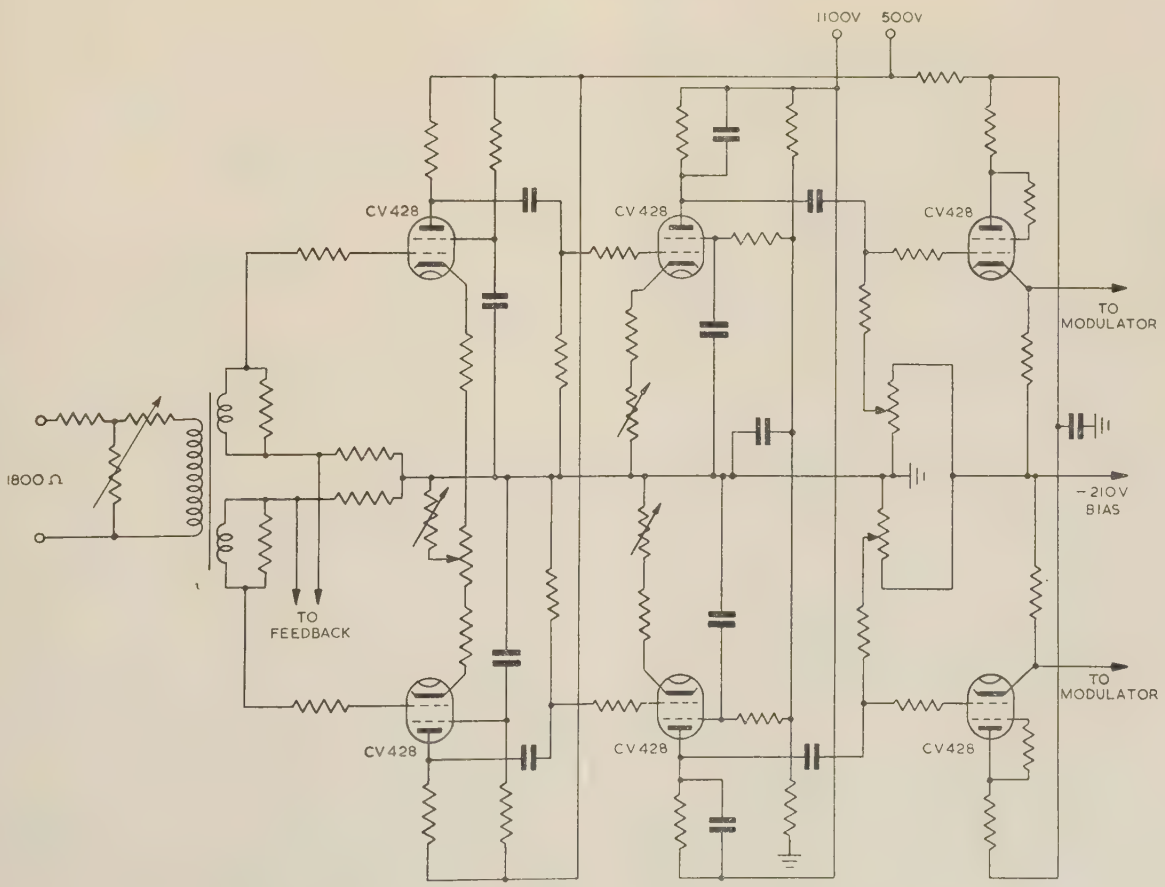


Fig. 7.—A.F. stages of modulator for 660-watt unit transmitter.

(3.3.1) **R.F. Stages.**

There are only two r.f. stages in this transmitter (Fig. 6). Drive is at radiated frequency to the grid of the first valve, which requires voltage drive only, so that isolation of one r.f. unit will not influence the magnitude of the drive to the remaining operating stages. The tuned circuit in the grid of the first stage (two CV428 tetrodes in parallel) is used to give the small degree of phase correction required by each transmitter, and the anode circuit is centre-tapped so that a voltage may be taken off the anti-phase point to neutralize the next stage—two DET27 triodes in parallel connected by a π-circuit to the output terminal. This circuit has shown itself to be particularly simple and reliable, and very stable in phase. Table 4 lists the characteristics of the valves.

Table 4

VALVE CHARACTERISTICS

	CV428	DET27
Heater voltage, volts	6·3	10
Heater current, amp	0·9	10
Anode voltage, volts	475	2 700
Anode dissipation, watts	16·5	250
Screen-grid voltage, volts	300	—
Screen-grid dissipation, watts	3	—
Mutual conductance, mA/volt	6*	9†
Amplification factor	9	22

* At $V_a = 300$ volts, $I_a = 72$ mA, $V_{g2} = 250$ volts.
† At $V_a = 700$ volts, $I_a = 150$ mA.

Large components of utmost mechanical simplicity were chosen for the circuits.

(3.3.2) **Modulator Stages.**

The modulator (Fig. 7) comprises three push-pull stages using six CV428 valves, driving the main class-B modulator employing two DET27 valves. The input impedance to each modulator stage is 1 800 ohms, so that three such stages in parallel will match the incoming a.f. line. The first stage is transformer-coupled, following an attenuator, to two *RC* coupled a.f. amplifiers in push-pull, and this is followed by a push-pull cathode-follower which drives the final modulator. Across the primary of the modulation transformer is an *RC* network forming the negative-feedback potentiometer, from which voltage is taken back to the grids of the first a.f. stage. The secondary of the modulation transformer is blocked from the d.c. line and modulates the final r.f. amplifier.

(3.3.3) **Power Supplies, etc.**

Each transmitter requires a single-phase mains supply and this permits the use of a 3-phase supply to three transmitters. The main h.t. supply to both the modulator and the final r.f. stage is at 1 600 volts and is derived from selenium-plate rectifiers. The auxiliary h.t. supply is at 600 volts and again is obtained from a selenium rectifier. Bias supplies also use similar rectifier elements.

A.C. filament heating is used. The built-in contactor circuits are completely interlocked, so that the transmitter is started by the closing of an external contact, usually of the time-switch type.

(4) PARALLELING ARRANGEMENTS

Earlier methods of paralleling the outputs of transmitters consisted of direct metallic connection between the outputs from the two transmitters, output coupling being adjusted to give the correct impedance to each of the two transmitters. This arrangement meant that the h.t. supply had to be applied simultaneously to the paralleled transmitters and also that a fault on one of them would affect the others.

For unattended use such an arrangement has serious drawbacks and a system was devised* whereby the paralleled transmitters would feed into a combining unit of such a characteristic that each could be quite independent of the others, as far as both control and overload circuits were concerned. The circuit devised is shown in Fig. 8(*a*). This is the arrangement for two

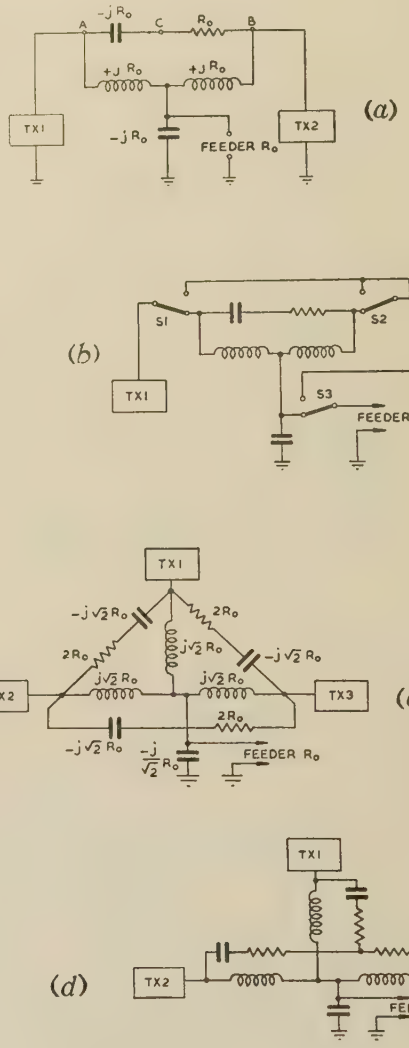


Fig. 8.—Combining circuits.

(*a*) Bridged-T for two transmitters.
(*b*) For two transmitters with by-pass switching.
(*c*) For three transmitters with delta ballast circuit.
(*d*) For three transmitters with star ballast circuit.

transmitters, and the essential feature is that it is non-dissipative in the normal condition and that voltage between point B and earth resulting from transmitter 1, and from point A to earth resulting from transmitter 2, is zero. Thus, whatever the condition of its fellow, each transmitter supplies its full power into its correct terminating impedance.

Unfortunately, although non-dissipative when the transmitters are correctly phased, such a network must have a dissipative element to ensure complete isolation as well as correct transmitter matching under all conditions, so that in the event of

* British Patent No. 743473.

the failure of one transmitter the signal strength would drop by 6 dB. Fig. 8(*b*) shows an arrangement of switches whereby, in the event of failure of one transmitter, the effective transmitter can be connected direct to the feeder by means of a momentary break in the h.t. supply. Switches S_1 and S_2 may be relay operated, so that the change can take place in 2 sec. Correct phase of transmitter outputs is indicated when the current in the ballast load resistor is zero.

Fig. 8(*c*) shows one possible arrangement when three transmitters are required to operate in parallel. This is essentially the same as that already described, inasmuch as the same conditions of transmitter isolation and loading apply. A rather better arrangement is shown in Fig. 8(*d*): this is simpler to engineer, since the circuit is more compact physically, and consequently the stray capacitances which reduce the effective isolation can be kept to a minimum. With the 3-transmitter combination the field strength of the radiated signal is proportional to the number of transmitters in service.

The use of an automatically controlled relay to by-pass sections of the combining unit in the event of such failure was considered to introduce a further possibility of hazard. Consequently these were not used and the combining circuit—the only common part of the equipment—was made particularly robust.

(5) ACKNOWLEDGMENT

The authors desire to thank the Engineer-in-Chief of Marconi's Wireless Telegraph Company, Limited, for permission to publish the paper.

DISCUSSION ON THE ABOVE TWO PAPERS BEFORE THE RADIO AND TELECOMMUNICATION SECTION, 10TH APRIL, 1957

Sir Harold Bishop: The B.B.C. began work on the possibility of reducing monitoring staff in 1947, and a paper* on the subject was read before The Institution in 1950. Mr. H. B. Rantzen did much of the early work on this subject, which was later extended to the control of transmitters, and the B.B.C. has now saved some 200 men—a salary saving of well over £100 000 a year.

I should like to have seen in the paper a clearer justification for the very considerable design work and capital and running costs of providing and maintaining the equipment set against the staff savings, improved performance and other advantages of the scheme. Other countries have similar automatic arrangements, but, generally speaking, they are not so complicated as those adopted by the B.B.C. but are confined to simple indications of transmitter conditions, modulation level, etc. I think it could therefore be argued that the B.B.C. arrangements err a little towards excessive complication and cost.

At the Commonwealth Broadcasting Conference in Australia at the end of 1956 it was reported that the broadcasting organizations of Australia, South Africa and Canada were operating unattended stations. In Canada, low-power unattended transmitters are located near railway stations with aerials not more than 100 ft from railway power and telephone lines. Reradiation from these lines provides useful reception for listeners living near the railway at considerable distances from the transmitter. The transmitters are extremely simple in design, and are often maintained by personnel with little technical knowledge. In South Africa the power of the transmitters is usually 250 watts, and the transmitters are never switched off; in this way, component failures are apparently reduced. Similar arrangements exist in Australia. In all these countries, as in the United Kingdom, most of the failures are due to breaks in power supplies and in programme lines.

I should like the authors' comments on the subject of reducing breakdowns by reducing maintenance and keeping the transmitters switched on. There is a tendency now in many countries to adopt this course, and there seems little doubt that with the increasing reliability of modern components, there are fewer interruptions to service if detailed preventive maintenance is discontinued and components disturbed as little as possible. For example, it is now common practice never to remove valves from sockets. I was told in Australia that the Australian Post Office has reduced maintenance work by 20% in this way.

As the authors have said, it is important to extend the unattended technique to television transmitters. The B.B.C. hopes to build a number of low-power unattended television transmitters, and we can look forward to considerable staff savings by adopting this technique.

With regard to the paper by Messrs. Morcom and Bowers, I should like to say that the transmitters at Daventry, and at the low-power stations to which reference has been made, have worked excellently, and I have nothing but praise for their performance. The feeling in the B.B.C. is that the paralleled-unit transmitter is better than the 'one on–one off' method. Finally, I should like to pay tribute to the manufacturers and the Post Office for their co-operation in the development of this work; we have had to rely on them a great deal. Moreover, the B.B.C. could not have carried through these major operational changes without the help and co-operation of the Association of Broadcasting Staff, who have always been ready to try out new ideas which in all cases have proved to be to the ultimate benefit of the staff concerned.

Mr. B. N. MacLarty: The possibility of using unattended transmitters was first discussed in 1947, and at that time, while one was convinced that reliable low-power unattended transmitters could be used, the idea of applying the principle to 100 kW units did not seem to be quite so attractive. I think this was because, in those days, all transmitters of this power were water cooled: water-cooling systems were apt to be troublesome and would introduce an element of unreliability in a completely unattended station. However, about this time, air-cooled valves of sufficient capacity became available and it was therefore possible to eliminate the principal element of unreliability.

The principle of using transmitters in parallel was pioneered by the B.B.C. The first high-power transmitters—two 200 kW units—were paralleled in about 1941. This experiment proved to be entirely successful, and the principle has now been extended to include television transmitters operating at 200 Mc/s

Sir Harold Bishop suggests that the monitoring equipment described in the papers is too elaborate and is rather attracted to the idea of using transmitters permanently switched on. I believe that it is now possible to design a transmitter which could be switched on and left unattended for perhaps a year. Such a transmitter would, of course, be somewhat more expensive to build, but having regard to the fact that components can be designed to have an almost indefinite life and the valves now in use have an operating life of over a year, such a development seems to be possible. It would then be possible to simplify the monitoring equipment considerably. If it is possible to lay across the Atlantic a cable with a great number of repeaters, which must run unattended for many years, it is difficult to see why the same principle cannot be applied to broadcasting transmitters.

* RANTZEN, H. B., PEACHEY, F. A., and GUNN-RUSSELL, C.: 'The Automatic Monitoring of Broadcast Programmes', *Proceedings I.E.E.*, Paper No. 1045 R, August, 1950 (98, Part III, p. 329).

Capt. C. F. Booth: The papers are representative of the present-day trend in both radio and line communication system design, namely increased reliability with reduced staff costs, which we have seen develop in both our cable (land and sea) and long-distance radio networks. Examples are the transatlantic telephone cable and the new transmitting station at Rugby. I am gratified to have confirmation of the considerable progress which has been achieved by the B.B.C. and industry in the broadcasting field; the authors and their colleagues have done a first-class job. I have no doubt that the next ten years will see even greater advances in the unattended-station technique for radio transmitting and receiving stations and cable repeater stations.

Referring to the B.B.C. paper and the question of reliability, I have one or two points on which I should like clarification. Will Messrs. Wynn and Peachey indicate the percentage of the transmission time for which the transmitters covered by Fig. 8 were on reduced power? I assume that the very low transmission-time loss due to valve failures (0·0001%) was due in part to the exclusion of those periods on which the sets were operating on reduced power due to valve failure. If the reference to a shut-down of 0·019% due to local failures at fully attended transmitters is directly comparable with that of 0·0093% for semi-attended stations, it seems clear that the present maintenance procedure at the former introduces incipient faults.

My final point concerns the application of the unattended-station technique to television transmitters. In my view this will assume ever-increasing importance; it will be used for small 'booster' stations operating in Band I, will be more useful for such stations in Band III, and will be very necessary when we come to exploit Bands IV and V. National coverage by these higher bands will involve a relatively large number of stations so that unattended stations will be an economic necessity. I therefore hope that the authors will continue to direct their efforts to this end.

Dr. A. J. Biggs: I am particularly glad to hear that automatic control is likely to extend to television transmitters. It seems to me that the television problem may be not much more difficult than the sound problem, because, after all, the transmitter has to accept a waveform and then radiate it at another frequency at a higher power. Indeed, perhaps with television a little more waveform distortion could be tolerated than is the case with sound. I should like to hear the views of the authors on this point.

It was interesting to note that components seemed to cause more trouble than valves, and I wonder whether the work has suggested any fundamental lines of investigation which should be pursued.

With regard to the electricity-supply figures given, I wonder whether the time lost was due to the decreasing reliability of supply mains or to the increasing use which the B.B.C. may be making of the supply mains in much of its equipment.

Mr. A. C. Beck: A critical appraisal of the benefits derived from the use of automatic monitoring equipment is difficult, in view of the number of factors involved. Although it was expected that the monitoring equipment would at times shut down workable transmitters, thus decreasing the overall reliability, this state of affairs is not reflected in the 0·019% quoted for the reliability of the fully attended station compared with the 0·013% for the unattended low-power stations.

It seems a pity that it is impracticable to provide duplicate aerials on medium waves as a means of increasing reliability, because feeders and aerials are known to fail, to the detriment of the reliability records of one hour per annum. The latest semi-attended 300 kW m.w. transmitter at Stockholm uses a duplicate aerial and is the first example of a split aerial fed by two equal-power transmitters with a maintained phase difference between aerial currents. Do Messrs. Wynn and Peachey consider it desirable to improve still further the reliability of transmitters, in view of the predominance of power and line failures?

It is gratifying to note that Messrs. Morcom and Bowers chose class-B modulation for the semi-attended transmitter, and it is interesting to recall that the first 100 kW m.w. transmitter supplied to the B.B.C. to use this system, namely Stagshaw, celebrates its 20th anniversary this year. While the transmitter described is an example of good engineering practice, do the authors consider that this design is suitable for future semi-attended stations? Is it not desirable to reduce the number of valves in the equipment and still further improve valve life? Would it not be desirable to eliminate the overall r.f. feedback and hum-cancellation equipment? As regards the high-power transmitters, is it not a fact that air-blast cooling is too cumbersome and too expensive in building space? For future transmitters, is it not desirable to introduce evaporation-type cooling, thus eliminating all rotating machinery and opening the way to the recovery and utilization of the waste heat?

Mr. M. D. Stonehouse: As a public supply engineer, my principal interest is the figures on power-supply failures given by Messrs. Wynn and Peachey. Having had access to more detailed information than appears in the paper, I must confirm that the figures given are, practically speaking, right, although one or two quite interesting things came out of the investigation. All the stations dealt with in the paper are, in effect, the B.B.C.'s reply to their 'rural area' problem: consequently, they rely upon the supply industry to provide their power in what is, in effect, a rural area so far as power supplies are concerned. Many of the stations are fed by 11 kV lines, and some of them are about 20 miles from the nearest supply point and are subject to all the winds that blow in heaven and to anything else that falls out of it. On this basis the figures given do not show a bad performance at all, although I have no doubt that they will be improved. Nothing will give us greater pleasure than to see the time lost due to power failure reduced to zero.

Mr. J. Rhodes: The remote and automatic control of broadcasting transmitters has made very exacting demands on the land lines connecting them to the programme source. The Post Office makes every endeavour to provide the lines required by the B.B.C., and regular meetings are held between the engineers of the two organizations to try to improve their performance further.

I would draw attention to the long distances over which these transmitters are controlled—100, 150 and even 250 miles. Many routes were not planned to provide a high-quality broadcast circuit, and we have to make various non-standard arrangements in order to provide the lines. For instance, we may have to amplify at ten or more repeater stations, and, unlike the transmitter, each repeater station must be equipped with standby power plant.

It appears that the programme time lost due to the transmitter is about 0·01%, that due to the line is 0·05% and that due to power failure is 0·06%. It should be noted that the loss due to the transmitter is with paralleled transmitters and is presumably for complete loss of programme. Duplicate or parallel land-lines are not provided over these long distances, but if we were able to do so, on a simple probability basis the time lost due to line faults would be of the order of 0·000 002 5%. I am not saying that this is a very accurate statistical calculation, and other factors may apply, but even if I am 100 times in error, the performance of the line would still be 40 times better than that of the transmitter.

We are not completely happy about the use of the earth phantom circuits, for the system is liable to a certain amount of interruption from earth currents; moreover, we have recently

traced one or two obscure faults to magnetic storms. I am not suggesting that the line engineer will adopt the practice of using the sunspot and its associated magnetic storm as an alibi for all his troubles: we shall still continue in our investigations, to give the best possible service.

Sir Noel Ashbridge: The figure of 0·06% for the time loss due to power failures seems to me rather surprising. I cannot remember the exact figures that we used to get from the smaller B.B.C. transmitters before the war, but I think that they averaged about 0·02% over several years. Does this mean that the power supplies before the war were much better than at present, or is there something wrong with the figure given? It seems to be very high indeed, and I wonder whether some very big breakdown is included in it. If you get a failure for an hour or two, due to fire perhaps, it affects the figure seriously for the whole year.

Col. J. D. Parker: Since I have had some experience of using radio relay for feeding the programmes to the transmitters, I wonder whether the authors can give any information on future policy regarding this aspect. If the concept of the network, whether medium wave, v.h.f. or television, is worked out on the basis of radio feed from a mother station, my experience of this type of operation is that the reliability factors can be of at least the same order as those of the present transmitter equipment.

To improve the land-line situation there is already to hand the radio rebroadcasting technique; this offers the equivalent of parallel operation of the lines, and it is indeed possible to achieve treble operation of the programme-feed circuits. I know that there are difficulties regarding finance and that the Post Office must watch this question carefully. Often a line must be provided for standby to the radio, and, therefore, the economics of the question are not easy. However, this method of programme feed has been used extensively in Europe, and I should like to know what the authors feel about future policy with medium-wave, v.h.f. and television programme feeds, possibly using the frequency-changer types of transmitter.

Mr. J. M. Chorley: The authors mention the telephone indicator panel which sends information back from the transmitter to a staffed centre. Have they in mind any device which could do the reverse, and send information back to the transmitter to get some action initiated, e.g. to extend the transmission hours beyond the normal time?

Have the authors any device in mind for countering the trouble which they may have experienced in the use of earth-phantom circuits as monitoring circuits?

THE AUTHORS' REPLIES TO THE ABOVE DISCUSSION

Messrs. R. T. B. Wynn and F. A. Peachey (*in reply*): Sir Harold Bishop seeks a justification for the initial cost of monitoring, and to that end an estimate has been made of the cost of apparatus, installation, etc., of all this automatic equipment. The capital cost is between £50000 and £100000 and the saving exceeds £100000 per annum. This initial expenditure could have been considerably less if the B.B.C. had been prepared to lower its standards of quality. Quality of service is measured, not only by the fidelity of frequency response, etc., but primarily by continuity of the correct transmission, and while land lines are used to serve these stations, a watch on the transmission is necessary.

With the present equipment and broadcasting hours we think there is a slight gain in reliability by switching off apparatus when it is not required. We certainly agree that it is a good policy to leave apparatus alone, and the transmitter and monitoring apparatus is designed so that it requires little or no servicing until it manifests its own faults.

Mr. MacLarty is doubtless correct in assuming that the simplicity introduced by valve air-cooling systems has helped the high-power unattended-station technique. We agree that some modern transmitters are so reliable that we can practically dispense with monitoring, and we rely on the normal failure indications commonly provided on such equipment. Circumstances have necessitated that the transatlantic telephone cable should be engineered to a standard of reliability far higher than is applied to normal broadcasting circuits, and the users of all such 2-way circuits provide the equivalent to automatic monitors.

Captain Booth is correct in assuming that valve failures which do not cause a complete shut-down of the station are not included.

The figures for periods of reduced radiated power are not available for these small stations, but it is not a significant trouble from the listener's point of view. The main reason why local faults on attended stations are slightly higher than at semi-attended stations is that at the former the reserve transmitters are neither running in parallel nor are they powered until breakdown occurs.

Both Captain Booth and Dr. Biggs express interest in the proposed unattended television stations. At first these will be of low power and almost certainly of the frequency-translator type, in which there is no great difficulty in avoiding waveform distortion.

Mr. Beck is certainly correct in assuming that, in some ways, improvement to the present transmitters is hardly worth while until we can attain improvement in the reliability of both power supplies and the means of feeding programme to them. On the other hand, the presence of staff in the vicinity of the transmitters means that there is often someone who can be summoned almost immediately when a fault is signalled. If transmitters were more reliable it might be possible to reduce this already meagre staff even further.

We shall be very pleased if Mr. Stonehouse can realize his hope that 'the time lost due to power failure becomes zero,' and we wish to express our appreciation of the attempts which are made to maintain our service.

We realize that the performance of earth phantom circuits is sometimes influenced by sunspots, but during the last several years these phantom circuits have served us well.

Sir Noel Ashbridge raises an interesting point about the apparent decrease in the reliability of power supplies, and we believe that this has two causes. The supply authorities have had difficulty in keeping pace with the demand, so that breakdowns are possibly more frequent now, but—as Mr. Stonehouse pointed out—we do tend to site our small unattended transmitters in remote areas and this tendency has certainly increased in the last few years. Moreover, before the war, some of our transmitters were powered by their own Diesel-generator sets or had them as standby units.

Regarding the remarks made by Col. Parker, we are, of course, rebroadcasting from radio receivers whenever it is practicable to do so. The new v.h.f. system is helping this in the Sound Services. The main difficulty, however, is that the received radio signal is usually weak at points where we wish to rebroadcast.

Finally, Mr. Chorley gives us the opportunity to say that, although we have in mind various new automatic devices of the kind he suggests, we are hesitant to add complication to the automatic control and monitoring system at a time when we hope that improvement in reliability will enable it to be simplified.

Messrs. W. J. Morcom and **D. F. Bowers** (*in reply*): Although high-power medium-frequency transmitters have been operated in parallel prior to the Daventry installation, we believe that this is the first time the advantage of the bridged-T scheme has been exploited at these powers and frequencies. The results have been very successful.

Captain Booth's remarks on unattended television stations emphasize the equipment designers' and, of course, the operators' problems. Straightforward frequency-changing repeaters filling local areas and more elaborate high-power stations capable of giving picture 'clean-up' before reradiation are both likely to be wanted in the future.

Dr. Biggs's remarks on permitted distortion can best be answered by saying the frequency-changing repeater, being the less complicated circuit, should be the more reliable. The specification of performance, however, lies with the operating authority.

The statements that components cause more trouble than valves needs a little explanation: it is true of certain types of component, usually those involving contacts, so that small relays and contactors must be used very conservatively, and any design features which eliminate their use contribute to overall reliability.

Mr. Beck states that the semi-attended 300 kW m.f. transmitter at Stockholm uses a split aerial fed by two equal-powered transmitters with a maintained phase relationship between aerial currents. This raises some interesting points. We believe that full benefit of paralleled transmitters is obtained only when they are isolated electrically in all respects. With economics in mind it is difficult to see how two m.f. aerials can be erected without mutual coupling. Thus, the two transmitters would be mutually coupled, and in the event of one failing the load imposed upon the other would change relative to that imposed when the transmitters were paralleled. The consequence of this change of load impedance is that a fault in one equipment is reflected into the other. Thus, both transmitters might be removed for a fault on one, and service would be interrupted.

Our policy is to maintain full electrical isolation between transmitters while maintaining their load impedances constant under all conditions. Thus, for m.f. equipments a lumped-circuit combining unit and single aerial system would be used. For higher frequencies (Bands I, II and III) a choice is available of paralleling via a network or the use of separate antennae and feeder systems.

On the question of choice of transmitter type for a semi-attended station we can make a few comments. First, the suitability of the transmitters at Daventry has been demonstrated by their performance record, and it is of interest to note that, when these transmitters were designed, there was no application for semi-attended use. The designers aimed at producing an equipment for competitive sale in world markets, and the choice of circuits, valves and other features was directed to this end. The general simplicity of the resulting design made it easy to adapt to semi-attended operation.

So far as future trends for semi-attended transmitters are concerned, Mr. Beck is quite right in stating the case for Class-B modulation. However, the improvements of recent years are considerable, and any new equipment would show savings in valves, power consumption, conversion efficiency and control simplification, when compared with older equipments.

Air cooling was criticized by Mr. Beck on the score of its being cumbersome and expensive in building space, but he neglected to mention reliability. While we agree that, to convey equal quantities of heat, an air duct is much larger than a steam or water pipe, we are firmly of the opinion that air cooling has distinct advantages. Air can be used directly through the valve radiator over a wide range of ambient temperatures (at least -30° C to $+95^\circ$ C) and the warmed outlet air can be used for space heating if desired. Modern air-cooled valves have light radiators and can be easily manhandled. The use of efficient valves and circuits does not require the high loss permitted by evaporation-type cooling, and the proper use of air cooling permits components other than valves to operate under better conditions. The biggest criticism of evaporation cooling is the fact that water is used. Only brief mention of the troubles caused by electrolysis, level control, steam-flow monitoring, radiators and corrosion is necessary. On the score of reliability we would rather pin our faith to the reliability of a rotating machine such as the 3-phase squirrel-cage motor driving a cooling fan, than to the elaborations that arise with any system using water.

DISCUSSION ON

V.H.F. BROADCASTING TECHNIQUE*

MERSEY AND NORTH WALES CENTRE, AT LIVERPOOL, 14TH JANUARY, 1957

Mr. A. J. Holliday: In view of the difficulties of maintaining accurate alignment in f.m. receivers, and the disadvantage of the inefficient limiting properties of the ratio detector, or the expense of a good discriminator, did the B.B.C. consider the possible merits of a pulse-width-modulated v.h.f. system? This system could be received satisfactorily, with less critical tuning of the receiver, demodulated by conventional means, and at the same time could share some of the advantages of frequency modulation.

Mr. D. R. Parsons: The v.h.f. scheme outlined by the authors is only a temporary solution to good sound reception. Restriction of the bandwidth of v.h.f. receivers to avoid adjacent-channel interference will undoubtedly occur, although the audio quality will not suffer, while manufacturers may tend to reduce the bandwidth of receivers in an effort to obtain more gain. This will mean that listeners will constantly have to adjust the tuning of their receivers in order to keep the carrier in the appropriate position of the i.f. bandwidth. Already many sets on the market are bad in this respect and cannot be left unattended unless crystal-controlled local oscillators are fitted or automatic frequency control is incorporated. If an f.m. receiver remains off-tune, man-made interference becomes extremely bad.

In December, 1955, a member of the B.B.C. staff wrote that 'during the months of May–July, 1955 (a period of exceptionally favourable conditions for tropospheric propagation), reception of Continental stations in the frequency range we are considering was observed at the B.B.C. Receiving Station, Tatsfield, on 46 days'. During the field tests conducted by the B.B.C., receivers were installed in the homes of members of the public and the B.B.C. staff. I presume that these sets were of the ideal communication standard of design and construction, and not the type supplied to the public to-day, in which a turn or so of wire is soldered across a wafer-switch contact as a cheap method of providing the inductance required to tune to the new frequencies.

In Section 3.3.1 it is stated that sound programmes are distributed by Post Office lines. How many of the inputs to the B.B.C. v.h.f. stations are limited in top l.f. response by these circuits? I should be very surprised if the l.f. response exceeded 8 or 9 kc/s.

The quality of the so-called 'high quality' v.h.f. receivers is limited by the miniature loudspeakers which are used, and so long as the B.B.C. cannot guarantee that the receivers do justice to their transmitters, it is invidious that a listener must buy an adaptor or another set and aerial to get better reception. In largely populated areas like Merseyside, it is far better to use wire broadcasting to receive the B.B.C. programmes direct from the station, and 160 000 people in this area prefer this method.

Mr. H. Martin: During the last war, enemy aircraft were prevented from using our radio transmitters for direction-finding by the placing of all stations carrying the same programme on one frequency. Is such a scheme possible using v.h.f. frequency modulation?

Does the choice of horizontal polarization make the design of an omni-directional aerial for mobile vehicles a difficult matter?

I notice from the map that Anglesey, Liverpool and Holme Moss lie in a straight line, with Liverpool equidistant from both transmitters; furthermore, both transmitters are of high power and are on adjacent channels, In view of the authors' statement in Section 3.1.7, is adjacent-channel interference a possibility in the Merseyside area, thus necessitating a dipole and reflector type of aerial to eliminate pick-up in the rear direction? The situation is complicated by the fact that many Liverpool residents are of Welsh origin and may wish to receive either transmitter.

Mr. K. L. Rao (*India: communicated*): In Fig. 1 on page 216, aerial heights have been indicated, but work in this direction is not mentioned in the paper. I suppose this has already been done in numerous v.h.f. field surveys in connection with television broadcast planning.

Presumably the need for maintaining an f.m. monitor of the counter type has not arisen at these stations for checking centre-frequency shift and peak deviation, the accuracy afforded by the units incorporated in the first two equipments being adequate. The null detector incorporated in the equipment affords the necessary check on the deviation meters, but no such counterpart appears to have been provided for centre-frequency check.

I suppose the new method of using the voltage regulator for power control does not create conditions in valves for degenerate operation at any stage; this has probably been taken care of in the choice of circuits and valves in the design stages.

The five 'obstacles' to the use of direct frequency modulation appear to have been well overcome. In fact, the Wrotham transmitter was using this method during trials for deciding on the system of modulation, either a.m. or f.m. The v.h.f. transmitters appear to have been designed after a decision on this issue was taken, but it is observed that the well-established direct system is still being used. I suppose there is a definite reason for reverting to this system even after the technique of F.M.Q. has been established.

I suppose the combining filters in the aerial circuits give the necessary cut off for frequencies beyond $f \pm 100$ kc/s. Considerations of bandwidth versus distortion would indicate distortion slightly greater than 1% in such a case. This is perhaps of no consequence, the overall distortion still being low.

The performance specifications shown in Section 2 on page 227, would indicate that the a.f. bandwidth is only from 30 c/s to 10 kc/s. Does this imply that sound frequencies are not extended up to 15 kc/s? In such a case, would the degree of f.m. advantage be greater in the overall a.f. response in the receiver?

Messrs. E. W. Hayes and H. Page (*in reply*): In reply to Mr. Holliday, a pulse-width-modulation system was considered for the v.h.f. service, but unless such a system employs a much wider bandwidth than the f.m. system, the signal/noise ratio is substantially worse. Further details were given in an article by Kirke.†

Provided that the receiver design problems mentioned by

* Hayes, E. W., and Page, H.: 'The B.B.C. Sound Broadcasting Service on Very High Frequencies', Paper No. 2258 R, November, 1956 (see **104 B**, p. 213).
Mortley, W. S.: 'Frequency-Modulated Quartz Oscillators for Broadcasting Equipment', Paper No. 2302 R, December, 1956 (see **104 B**, p. 239).
Only the first paper was discussed at Liverpool; Mr. Rao's contribution refers to both papers.

† Kirke, H. L.: 'The Application of Pulse Technique to Broadcasting', *B.B.C. Quarterly*, 1946, **1**, p. 139.

Mr. Parsons are fully understood, their solution need not necessarily involve great complication, and consequently expense. For instance, the tuning drift of many receivers in the popular-price class is, in our opinion, reasonably small. We hope that more effective limiting will be achieved without increasing the cost of receivers appreciably now that the problem has been ventilated. The planning standards described in Section 3.1.6 and 3.1.7 were based not on the performance of the experimental receiver used for the field tests, but on that of commercial receivers available at the time. Periods of abnormal propagation were also taken into account.

Almost all the lines used for locally-originated programmes have a bandwidth of approximately 15 kc/s. At present the bandwidth of the long lines used for the national programmes generally lies between 7·5 and 10 kc/s, but is greater in some cases. The aim of the B.B.C. is to make it possible for the listener to obtain high-quality reception, provided that suitable receiving equipment is used. It seems to us neither practicable nor desirable for the B.B.C. to guarantee the performance of receivers.

The question of direction-finding using v.h.f. transmissions does not come within the scope of the present paper; the scheme mentioned by Mr. Martin is theoretically possible, but there are obvious practical difficulties.

The design of an omni-directional horizontally-polarized receiving aerial for a vehicle does not present undue difficulty unless there are restrictions on its position on the vehicle. In practice, the polarization is greatly modified in the vicinity of the vehicle; a vertical aerial may therefore be used, if required, for combined a.m.–f.m. reception. In this case the use of horizontal rather than vertical polarization probably makes reception difficulties somewhat greater; but the ability of the f.m. signal to give satisfactory reception, even with a very low input level, means that horizontal polarization is not a serious drawback in practice.

Some changes have been made to the national coverage plan since the paper was published. It has now been decided that the permanent Anglesey station will employ a relatively low-power transmitter; it is therefore very unlikely to interfere with reception of Holme Moss in the Liverpool area. The Welsh Service will be received well from the station near Corwen, the power of which will be greater than is shown in the published paper. Since the spacing between the Holme Moss and Corwen carriers is 400 kc/s, mutual interference in the Liverpool area is unlikely.

In reply to Mr. Rao, the height of the transmitting aerials was based on field trials carried out in planning the television service.*

The test equipment provided at transmitting stations includes means for checking the peak deviation and also the centre frequency of the carrier.

The voltage regulator takes 30 sec to run up the voltages from zero to the normal working values; filament, standing bias and anode voltages are increased together. The application of drive is postponed for a further 10–15 sec, during the heating-up of the indirectly heated filaments of the anode-voltage rectifiers of the low-power stages. The transmitter is stable throughout the static period. To avoid loading the main rectifier valves before the cathodes are fully heated, the rectifier filaments are maintained at 80% of full voltage while the regulator is in the 'off' position.

* TAGHOLM, L. F., and ROSS, G. I.: 'The Selection and Testing of Sites for Television Transmitters in the United Kingdom', *Proceedings I.E.E.*, Paper No. 1262 R, April, 1952 (99 Part III A, p. 300).

Two methods of modulation (described in the companion papers) are used because the equipment was provided by different manufacturers. The combining filters have a high stop-band loss and a low pass-band loss over a bandwidth of ±100 kc/s. Non-linearity of the phase-frequency characteristic of these filters would cause distortion of the transmitted signal; but in practice this is small, and the distortion in the complete transmitting installation is only about 0·5%.

Although the response and distortion of the transmitting equipment is specified in the companion paper only up to 10 kc/s, the response is, in fact, substantially uniform up to 15 kc/s. The signal/noise ratio advantage of frequency modulation over amplitude modulation is not greatly affected by the upper audio-frequency limit, provided that this is at least 6 c/s. The contrary is sometimes stated in theoretical treatments, but this is because the unrealistic assumption is made that the ear has a uniform response over the audio-frequency band.

Mr. W. S. Mortley (*in reply*): If Mr. Rao's second comment refers to the F.M.Q. equipment, he appears to have missed the statement on page 244: 'The carrier frequency is adjusted by beating the fifth harmonic of the crystal against a reference crystal oscillator'. This is a low-temperature-coefficient crystal not thermally controlled. A more accurate check may be given by a separate monitor receiver designed by Mr. L. G. Kemp and his associates. This utilizes a thermally controlled reference crystal and a thermally controlled Foster–Seeley discriminator, and has an absolute stability of about 1 part in 10^6. A counter-type discriminator could have been used but is not necessary and is not necessarily more stable. More accurate checks of centre frequency may be made from the B.B.C. receiving station at Tatsfield.

Presumably by the five 'obstacles' to the use of direct frequency modulation, Mr. Rao is referring to my remarks in the discussion about the direct frequency modulation of a crystal oscillator. He is quite correct in his statement that the Wrotham transmitter was using this method during the f.m.-a.m. trials. The B.B.C. are using both direct frequency modulation of crystal oscillators and direct frequency modulation of *LC* oscillators. At present there are more of the crystal type in service. So far as I know there is no question of a reversion to the other type. The existence of the two types in the B.B.C. service arises out of B.B.C. policy in regard to the placing of contracts with more than one contractor.

The frequency cut-off beyond ±100 kc/s. is not as necessary as Mr. Rao suggests. The worst out-of-band sideband interference would be given by 75 kc/s deviation due to a 15 kc/s modulating frequency. The eighth sideband, at 105 kc/s from the carrier, would have an amplitude 34½ dB below that of the unmodulated carrier, and the 9th, 10th and 11th sidebands, at 120, 135 and 150 kc/s, would be 45, 69 and 82 dB below, respectively. Higher modulation frequencies are prevented from having any appreciable amplitude by the low-pass filter in the a.f. input circuit. Such modulation is most improbable, in any case.

In assessing the importance of this rate of cut-off, one should compare it with the selectivity curve of the best receiver commercially available, and it will become quite clear that the effect of these sidebands is negligible. The receiver characteristics enforce planning which gives protection against mutual interference.

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THERMAL TURNOVER IN GERMANIUM P–N JUNCTIONS

By A. W. MATZ, M.Sc.

(*The paper was first received* 18*th March, and in revised form* 9*th August,* 1957.)

SUMMARY

The static reverse characteristics of an alloyed germanium p–n junction are analysed, taking into account an avalanche multiplication factor. The condition for thermal stability is examined, the onset of thermal runaway is related to a thermal turnover phenomenon, and the existence of a negative-resistance region is predicted.

The existence of such thermal negative-resistance regions in germanium p–n junctions has been verified experimentally, and the results are analysed semi-quantitatively, showing fairly good agreement with the theory.

LIST OF PRINCIPAL SYMBOLS

$b = W_G/k$

$b' = \dfrac{W_G - \frac{1}{2}W_A}{k}$

$b'' = \dfrac{W_G - W_A}{2k}$

D_n = Diffusion constant for electrons.
D_p = Diffusion constant for holes.
e = Electronic charge.
g_i = Isothermal differential conductance.
g_{ir} = Reverse isothermal differential conductance.
g_s = Static differential conductance.
g_{sr} = Reverse static differential conductance.
I_s = Saturation (or bulk leakage) current.
I_{s1} = Saturation current at the turnover point.
I_L = Surface leakage current.
$I_L(V)$ = Voltage dependent part of surface leakage current.
I_c = Collector current.
I_e = Emitter current.
J_s = Saturation current density.
k = Boltzmann's constant.
M = Avalanche multiplication factor.
M_1 = Avalanche multiplication factor at turnover point.
n = Exponent of V/V_B occurring in M.
n_i = Electron (or hole) density in intrinsic material.
N_d = Density of donor impurities.
N_a = Density of acceptor impurities.
p_n = Density of holes in n-type material.
R = Slope of load line (Fig. 2).
S = Thermal resistance.
T = Absolute temperature.
T_a = Ambient temperature.
T_1 = Turnover temperature.
T_{jmax} = Maximum permissible junction temperature.
V = Applied voltage.
V_B = Breakdown voltage.
V_k = Common-emitter breakdown voltage.
V_1 = Turnover voltage.
V_c = Collector voltage.
V_{cc} = Battery voltage.
W_A = Minority carrier lifetime activation energy.
W_G = Energy gap between valence and conduction bands.
α_{N0} = Low-voltage common-base current amplification factor.
λ = Exponent occurring in I_L.
ϕ = Exponent occurring in I_s.
$\theta_1 = T_1 - T_a$ = Excess turnover temperature.
ρ = Resistivity.
τ_n = Life-time of electrons.
τ_p = Life-time of holes.
μ_p = Drift mobility of holes.

Written contributions on papers published without being read at meetings are invited for consideration with a view to publication.

Mr. Matz is with Hivac, Ltd.

(1) INTRODUCTION

In the past the peak reverse voltage of a semi-conductor diode has had to be determined empirically, since the way in which the breakdown voltage expected from a simple avalanche mechanism is modified by the onset of thermal runaway has not been completely understood. It is the purpose of the paper to establish analytically that the conditions for operational stability can be determined according to the two extremes, namely

(*a*) *Constant-voltage source.*—The maximum reverse voltage at which the diode can be stably operated in this case is V_1. For voltages greater than V_1 thermal runaway occurs, the high current passed causing the fusing of the diode. V_1 depends on the isothermal avalanche breakdown voltage, V_B, and on the reverse current, both surface leakage and bulk components being important.

(*b*) *Constant-current source.*—Here it is the power dissipation which is limited by the upper limit of diode junction temperature, depending purely on the metallurgical and mechanical structure, with stable operating points possible on a negative-resistance portion of the static characteristic.

Intermediate source resistances fall into the first or second of these categories, according to whether they are lower or higher than a certain *minimum stable resistance.*

In Section 3 the static reverse characteristic of an alloyed germanium p–n junction diode is considered, and the current, voltage and junction temperature at the turnover position are determined. The analysis takes into account an avalanche multiplication factor and the influence of trapping centres in so far as it affects the temperature dependence of minority-carrier life-time. The turnover theory presented is analogous to that outlined by Burgess[1] in his work on thermistors and point-contact germanium diodes, Newton's law of cooling being assumed throughout. The form of negative resistance obtained is shown to be short-circuit unstable, and a formula for the curvature at the turnover point is given.

In Section 4 an upper limit to the turnover temperature for a 'Shockley' (i.e. no surface leakage) germanium junction is calculated, and the variation of the position of such turnover points with ambient temperature and the other diode constants is demonstrated graphically, an interesting feature being that, although V_1 may be close to V_B (which is its upper bound) at about room temperature, V_1 will generally lie well below V_B at elevated ambient temperatures. This is in accordance with observations[2] of permissible operating points.

The deviation is even more marked when, in addition, surface leakage current is present, as is also the departure of T_1 from the value $T_a + \phi$, obtained when only bulk current is

considered.[3] A modification of the theory to include surface leakage is offered in Section 6.

(2) THE CURRENT/VOLTAGE CHARACTERISTIC

According to Shockley,[4] the current density through a perfect p–n junction, is given by

$$J = J_s\left(\exp\frac{eV}{kT} - 1\right) \quad . \quad . \quad . \quad . \quad . \quad (1)$$

This expression was derived on the basis of the following assumptions:

(*a*) The transition region is so narrow that negligible recombination takes place there.
(*b*) The density of minority carriers on either side of the junction is very much less than that of majority carriers (i.e. $p_n \ll n_n$ and $n_p \ll p_p$),
(*c*) The electric field is negligible on either side of the transition region, and
(*d*) Steady-state conditions have been reached.

J_s is given by

$$J_s = e\left(\sqrt{\frac{D_p}{\tau_p}}p_n + \sqrt{\frac{D_n}{\tau_n}}n_p\right) \quad . \quad . \quad . \quad . \quad (2)$$

If the analysis is restricted to the case where the p-side of the junction is of much higher conductivity than the n-side (i.e. $p_n \gg n_p$), the saturation current will be carried mainly by holes, and, to this order, will be given by

$$J_s \simeq (J_s)_p = e\sqrt{\frac{D_p}{\tau_p}}p_n \quad . \quad . \quad . \quad . \quad . \quad (3)$$

When the opposite case is true, eqn. (3) can be replaced by $J_s \simeq (J_s)_n$, and the analysis can be modified accordingly.

Eqn. (1) does not take avalanche multiplication into account, and it is therefore necessary to introduce the empirical expression[5]

$$M = \frac{1}{1 - \left(\frac{V}{V_B}\right)^n} \quad . \quad . \quad . \quad . \quad . \quad (4)$$

in which n is governed by the resistivities of the two sides of the junction. Thus, for the collector junction of a p–n–p alloyed junction transistor, n is about 3, while for n–p–n units n is approximately 6, as described by Miller.[5] Methods of determining V_B and n are described below.

The expression for the current is therefore written

$$I = MI_s\left(\exp\frac{eV}{kT} - 1\right) \quad . \quad . \quad . \quad . \quad . \quad (5)$$

(2.1) Isothermal Characteristics

The isothermal differential conductance is given by

$$g_i = \left(\frac{\partial I}{\partial V}\right)_T = \frac{e}{kT}(I + MI_s) + \frac{I}{M}\frac{\partial M}{\partial V} \quad . \quad . \quad (6)$$

At reverse voltages higher than 0·1 volt the exponential term in eqn. (5) becomes negligible in comparison with unity, and the current assumes the value $I = -MI_s$. Substituting this value into eqn. (6) and differentiating M as defined in eqn. (4) reduces eqn. (6) to

$$g_{ir} = -\frac{I_s}{V}nM(M-1) \quad . \quad . \quad . \quad . \quad (7)$$

When the reverse voltage is very much less than the breakdown voltage, $M \simeq 1$, but as V approaches V_B, M approaches infinity, so that the reverse-bias isothermal characteristics are almost completely flat at low voltages, but soften as the voltage approaches an appreciable proportion of the breakdown value, the slope being given by eqn. (7). Of course, this formula does not take into account surface leakage or surface breakdown.

(2.2) Steady-State Characteristics

Let it be assumed that under d.c. conditions the junction is allowed to attain thermal equilibrium at each point of the characteristic, so that the electrical power $P = IV$ is exactly balanced by the heat dissipated from the junction, and that Newton's law of cooling is valid, so that $dT/dP = S$, a constant. Then

$$T - T_a = SP \quad . \quad . \quad . \quad . \quad . \quad . \quad (8)$$

The thermal-stability condition is interpreted as being that for which eqns. (5) and (8) are compatible. Under stable conditions, therefore, eqns. (5) and (8), taken together, represent in parametric form a family of steady-state characteristics for different ambient temperatures.

The steady-state differential conductance is defined by $g_s = dI/dV$. If dT is eliminated from the equations,

$$dI = \left(\frac{\partial I}{\partial V}\right)_T dV + \left(\frac{\partial I}{\partial T}\right)_V dT \quad . \quad . \quad . \quad . \quad (9)$$

and

$$dT = S(IdV + VdI) \quad . \quad . \quad . \quad . \quad . \quad . \quad (10)$$

g_s is obtained as

$$g_s = \frac{\left(\frac{\partial I}{\partial V}\right)_T + SI\left(\frac{\partial I}{\partial T}\right)_V}{1 - SV\left(\frac{\partial I}{\partial T}\right)_V} \quad . \quad . \quad . \quad . \quad (11)$$

Differentiation of eqn. (5) yields

$$\left(\frac{\partial I}{\partial T}\right)_V = \frac{I}{MI_s}\frac{\partial(MI_s)}{\partial T} - \frac{eV}{kT^2}(I + MI_s) \quad . \quad . \quad (12)$$

while $\left(\frac{\partial I}{\partial V}\right)_T$ is given by eqn. (6).

(3) ANALYSIS OF REVERSE-BIASED STEADY-STATE CHARACTERISTICS

On substitution from eqns. (6) and (12), and replacement of I by $-MI_s$, eqn. (11) reduces to

$$g_{sr} = \frac{-MI_s\left[\frac{n}{V}(M-1) - S\frac{\partial(MI_s)}{\partial T}\right]}{1 + SV\frac{\partial(MI_s)}{\partial T}} \quad . \quad . \quad (13)$$

(3.1) Thermal Stability

The equations

$$I = -MI_s \quad . \quad . \quad . \quad . \quad . \quad . \quad (14)$$

and

$$T - T_a = SIV \quad . \quad . \quad . \quad . \quad . \quad . \quad (15)$$

can conveniently be regarded as a set of current/temperature characteristics with voltage as parameter (Fig. 1).

At a specified voltage, V, eqn. (15) represents a straight line through the point $(0, T_a)$. The intersection of this with the curve of eqn. (14) gives the operative point of the diode, like P_1 or P_2. Each such point implies one possible power only, since eqns. (14) and (15) taken together admit of only one degree of freedom. At junction voltages which are low compared with the breakdown voltage, eqn. (14) gives I as essentially a function of temperature only; so that the corresponding operating points lie on the intersection of a family of lines from $(0, T_a)$ towards a static curve. For higher voltages the curve will move to the right as the slope of the lines increases.

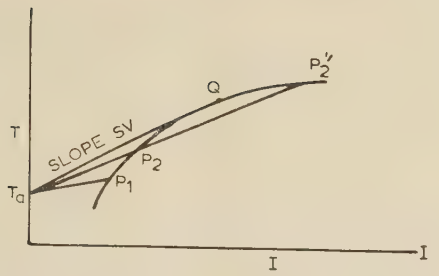


Fig. 1.—Compatibility of eqns. (14) and (15).

The position of limiting stability (I_1, T_1) occurs when such a straight line becomes a tangent to the curve at Q, i.e. when

$$\frac{I_1}{T_1 - T_a} = \left[\left(\frac{\partial I}{\partial T}\right)_V\right]_Q \quad . \quad . \quad . \quad . \quad (16)$$

which can be written

$$\frac{1}{T_1 - T_a} = \frac{1}{MI_s}\frac{\partial(MI_s)}{\partial T} \quad . \quad . \quad . \quad . \quad (17)$$

Theoretically, for a given ambient temperature, T_1 can be found from eqn. (17). Eqns. (14) and (15) can then be used to find the corresponding values of I_1 and V_1. Clearly, at voltages higher than V_1, instability will occur, and V_1 represents the runaway limit for a constant-voltage source.

(3.2) The Turnover Condition

Let it be assumed that for any given ambient temperature the steady-state curves contain a voltage maximum, V_1, at junction temperature T_1 and current I_1, at which point the onset of thermal runaway occurs if a constant-voltage supply is used. Then

$$1 = V_1\left(\frac{\partial I}{\partial T}\right)_{\substack{V=V_1\\T=T_1}}\frac{dT}{dP} \quad . \quad . \quad . \quad . \quad (18)$$

as given by Burgess.[1] For completeness, it can be verified that, when the steady-state conditions represented by eqn. (8) exist, eqn. (18) corresponds to the denominator of eqn. (13) vanishing, from which it can be deduced that I_1 and T_1 as defined by eqn. (16) are identical with turnover current and temperature respectively. This could easily be inferred from the fact that the voltage, which is proportional to the slope of the line family in Fig. 1, passes through a stationary value at tangency.

However, the current can be increased beyond I_1 to a point like P_2', thus reaching a negative-resistance portion of the characteristic. This increase in current is not compensated by a corresponding reduction in voltage, as the temperature rise indicates. Thus, the turnover point should be thought of as corresponding strictly to a voltage maximum rather than a power maximum, since stability is possible at higher powers than the turnover power. This appears to have been insufficiently stressed so far, and the temperature rise to cause thermal instability usually given[3, 6, 7] surely refers to the turnover point, shown as Q in Fig. 2 (see Section 3.3 for the series resistance required to obtain permissible stable operating points at higher temperatures).

(3.3) Form of Negative Resistance

The steady-state curves are seen to contain a voltage maximum (at the turnover point) but no minimum, and approach the asymptote $V = 0$ as $I \to \infty$. This portion of the curve is never observed in practice, since the diode would reach too high a temperature. The negative resistance is short-circuit unstable. The usual condition[8] for stability

$$R + r > 0 \quad . \quad . \quad . \quad . \quad . \quad . \quad (19)$$

where r is the slope resistance of the characteristic and R is the resistance of the associated load line, holds. The impedance of the d.c. generator used in the experiments described was kept higher than the negative slope resistance at the point of inflection (called the minimum stable resistance) occurring on all steady-state curves, and thus runaway was avoided.

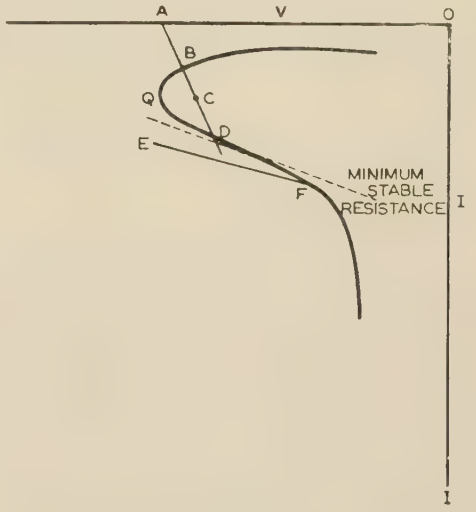


Fig. 2A.—Typical steady-state characteristic and position of load lines.

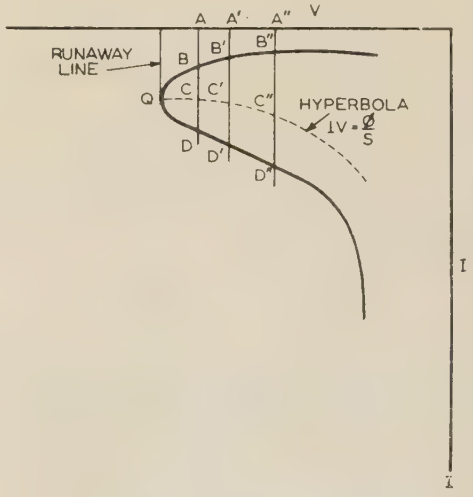


Fig. 2B.—Locus of stability demarcation points along constant-voltage-source load lines.

Fig. 2A shows a typical steady-state reverse characteristic, and AD, EF, A′D′, etc., denote load lines. Intersections such as B and F are stable, while those such as D are unstable. If the diode is suddenly switched to a point between A and B (e.g. from operation in the forward direction) it will generally become warm and settle at B; also for any point C between B and D, it will cool to B or heat according as

$$IV - I^2R \lesseqgtr K \quad . \quad . \quad . \quad . \quad . \quad . \quad (20)$$

The value of K is derived in Section 12, where it is also shown that for points like B and D, the conditions (19) and (20) are

equivalent. If the diode becomes heated it cannot settle at D, since this point is unstable, but could theoretically reach a stable position where the load line intersects the curve for the third time. In the majority of cases, however, this would give rise to a current corresponding to $T > T_{jmax}$, so that inequality (20) corresponds to the runaway condition for most practical purposes.

Inequality (20) separates stable from unstable regions along a load line. In Fig. 2B the locus of these demarcation points, C, for a set of load lines approximating a constant-voltage source is determined. The broken line in Fig. 2A is the tangent drawn through the point of zero curvature, the slope of which represents the minimum series resistance required to obtain stability at an operating point on the negative-resistance portion of the characteristic.

(3.4) The Solution for the Turnover Point

Expression (17) can be written

$$\frac{1}{T_1 - T_a} = \frac{1}{I_{s1}}\left(\frac{\partial I_s}{\partial T}\right)_1 + \frac{1}{M_1}\left(\frac{\partial M}{\partial T}\right)_1 \quad . \quad . \quad (21)$$

in which

$$\frac{1}{M}\frac{\partial M}{\partial T} = (M-1)\left[\frac{\partial n}{\partial T}\log_\varepsilon\left(\frac{V}{V_B}\right) - n\frac{\partial}{\partial T}\log_\varepsilon(V_B)\right] \quad (22)$$

is obtained by logarithmic differentiation of eqn. (4).

(3.4.1) Form of Saturation Current, I_s.

Let it be assumed that the approximation (3) is valid, and that[9]

$$p_n = \frac{N_d - N_a}{2}\left\{-1 + \sqrt{\left[1 + \frac{4n_i^2}{(N_d - N_a)^2}\right]}\right\} \quad . \quad (23)$$

Clearly,

$$\left.\begin{aligned} &p_n \to n_i \text{ for } n_i^2 \gg (N_d - N_a)^2 \\ \text{but} \quad &p_n \to \frac{n_i^2}{N_d - N_a} \text{ for } n_i^2 \ll (N_d - N_a)^2 \end{aligned}\right\} \quad . \quad . \quad (24)$$

corresponding to the intrinsic and extrinsic temperature ranges respectively. The dependence of n_i on temperature for germanium can be expressed by [10]

$$n_i(T) \propto T^{3/2}\exp\left(\frac{-W_G}{2kT}\right) \quad . \quad . \quad . \quad . \quad (25)$$

where $W_G \simeq 0{\cdot}7\,\text{eV}$. Combining expression (25) with the second approximation (24), since mainly the extrinsic range is relevant in the present study, one finds

$$p_n(T) \propto T^3\exp\left(\frac{-b}{T}\right) \quad . \quad . \quad . \quad . \quad (26)$$

The temperature dependence of D_p follows that of minority-carrier mobility by the Einstein relationship $D_p = \frac{kT}{e}\mu_p$, and, over the range of impurity concentrations usually considered, mobility has been found[11, 12] to vary approximately according to $\mu_p \propto T^{-2{\cdot}3}$. Thus

$$D_p(T) \propto T^{-1{\cdot}3} \quad . \quad . \quad . \quad . \quad . \quad (27)$$

The dependence of life-time on temperature in germanium p–n junction diodes has been thoroughly investigated by Pell.[13] At about room temperature, the relationship was found to be

$$\tau_p(T) \propto \exp\left(\frac{-W_A}{kT}\right) \quad . \quad . \quad . \quad . \quad (28)$$

with activation energies of about 0·3 eV. Since Pell's diodes exhibited no change in effective life-time when subjected to different surface treatments, it is suggested that the life-time referred to was essentially the bulk life-time, and the approximate figure for W_A of 0·3 eV is substantially in agreement with those recently obtained by Ransom and Rose[14, 15] concerning such life-times.

Although Webster[16] has reasoned that the effective life-time to be associated with alloyed germanium junctions is dominantly the surface life-time, no one appears to have considered the temperature dependence of surface life-time in completed junction devices. Consequently, the analysis below is valid only for bulk saturation current, surface effects, including surface breakdown, being neglected; a modification of the theory to include surface leakage current I_L, however, is offered in Section 6, where an empirical form for I_L is related to the experimental turnover curves.

Pell[13] also discovered that the reverse saturation current consists of two components, namely the usual diffusion term of eqn. (3) and a term dependent on charge generation by recombination centres in the space-charge region, although the latter was found to be important only at temperatures lower than considered here. Thus, by substituting expressions (26)–(28) into eqn. (3), the saturation current is written as

$$I_s(T) \propto T^m\exp\left(\frac{-b'}{T}\right) \quad . \quad . \quad . \quad . \quad (29)$$

where $m \simeq 2{\cdot}35$ and $b' = (W_G - \frac{1}{2}W_A)/k$.

Differentiating eqn. (29) logarithmically gives

$$\frac{1}{I_s}\frac{\partial I_s}{\partial T} = \frac{mT + b'}{T^2} \quad . \quad . \quad . \quad . \quad . \quad (30)$$

and eqn. (21) becomes

$$\frac{(1-m)T^2 + (mT_a - b')T + b'T_a}{T^2(T - T_a)} = \frac{1}{M}\frac{\partial M}{\partial T} \quad . \quad (31)$$

(3.4.2) Neglect of $\frac{1}{M}\frac{\partial M}{\partial T}$.

It is clear from expression (22) that since

$$\lim_{V \to 0}(M-1)\log_\varepsilon\left(\frac{V}{V_B}\right) = 0$$

it may be possible to neglect $\frac{1}{M}\frac{\partial M}{\partial T}$ for turnover voltages well below breakdown, since $\partial n/\partial T$ is unlikely to be very large. However, at turnover voltages close to V_B, with $M \gg 1$,

$$\frac{1}{M}\frac{\partial M}{\partial T} \simeq -Mn\frac{\partial}{\partial T}(\log_\varepsilon V_B) \quad . \quad . \quad . \quad (32)$$

It has been found that, for germanium junctions, there is very little change of V_B with T at temperatures less than some 60° C or so above room temperature.[17] As shown below, it is the lower ambient temperatures which correspond to $V_1 \to V_B$, so that under conditions in which eqn. (32) prevails, again it may be possible to neglect $\frac{1}{M}\frac{\partial M}{\partial T}$, even though M is large.

Neglecting the right-hand side of eqn. (31), therefore, and letting the excess temperature above ambient be θ_1, it is found that θ_1 is the positive root of the quadratic

$$(m-1)\theta^2 + [b' + (m-2)T_a]\theta - T_a^2 = 0 \quad . \quad (33)$$

i.e. $$T_1 - T_a = \theta_1 = \frac{b' + (m-2)T_a}{2(m-1)} \left(-1 + \sqrt{\left\{1 + \frac{4(m-1)T_a^2}{[b' + (m-2)T_a]^2}\right\}} \right) \simeq \frac{T_a^2}{b' + (m-2)T_a} \quad . \quad (34)$$

for $b' \gg T_a$, which is usually the case. For $T_a = 300°\,\mathrm{K}$, $b = 8\,700°\,\mathrm{K}$ (germanium), $W_A = 0{\cdot}3\,\mathrm{eV}$ and $m = 2{\cdot}35$, it is found that $\theta_1 \simeq 14°\,\mathrm{C}$.

Also, $$\frac{d\theta_1}{dT_a} = \frac{2T_a - \theta_1(m-2)}{2(m-1)\theta_1 + (m-2)T_a + b'} \quad . \quad . \quad (35)$$

obtained by differentiation of eqn. (33), is clearly positive but small. Thus, if only ambient temperatures greater than 300° K are considered, the value of θ_1 will be at least 14° C for germanium junctions, but not very much more.*

If the right-hand side of eqn. (31) is different from zero, then according as the sign of $\partial M/\partial T$ is positive or negative, the solution (34) represents an upper or lower limit to T_1. When eqn. (32) holds, the value of $\partial M/\partial T$ depends on the temperature coefficient of V_B. For silicon diodes, Pearson and Sawyer[18] found a linear relationship $V_B(T) = V_B(T_a)[1 + \beta(T - T_a)]$, where $V_B(T_a)$ represents the room-temperature breakdown voltage, and β was found to be positive for the step junctions considered. McKay[19] also found that β was positive for silicon graded junctions, but found $\partial M/\partial T$ to be relatively unimportant. For germanium junctions, however, only reductions in V_B with T have been reported.[17] This may be due to deterioration of the surface, leading to surface breakdown at a voltage less than V_B. Until experiments can be performed to select those units which exhibit volume rather than surface breakdown, this uncertainty will remain. Thus, there would appear to be no likelihood of the junction temperature, T_1, at turnover exceeding the calculated bound at ambient temperatures low enough to ensure the validity of eqn. (32).

In the following numerical analysis, therefore, $\frac{1}{M}\frac{\partial M}{\partial T}$ is neglected on the assumptions previously stated, and the T_1 values calculated refer to an upper limit to turnover temperature, if not giving the actual values.

(3.4.3) **Value of Turnover Voltage.**

It is clear from eqns. (14) and (15) that the turnover voltage is given by

$$V_1 = \frac{\theta_1}{SM_1I_{S1}} \quad . \quad . \quad . \quad . \quad . \quad . \quad . \quad (36)$$

in which $M_1 = \dfrac{1}{1 - \left(\dfrac{V_1}{V_B}\right)^n}$ is the multiplication factor at the turnover point and I_{s1} is $I_s(T_1)$. The turnover power, $P_1 = \theta_1/S$, is approximately constant, as shown by eqn. (35).

With the example just considered, namely $T_a = 300°\,\mathrm{K}$ and $\theta_1 = 14°\,\mathrm{C}$, and with typical values of $S = 0{\cdot}35°\,\mathrm{C/mW}$, $I_s(300°\,\mathrm{K}) = 4\,\mu\mathrm{A}$, say, it is seen from eqns. (36) and (29) that M_1 must be large, so that $V_1 \simeq V_B$. It is inferred from this that the turnover voltage, whilst always being bounded above by V_B, coincides with it more closely at low ambient temperatures (e.g. room temperature), but at higher values of T_a, I_1 increases and V_1 falls and begins to depart from V_B.

For $n = 3$, V_1 departs by 5% from V_B when θ_1/SI_{s1} reaches $6{\cdot}65V_B$. From eqn. (35) it is known that θ_1 is more or less fixed, so that for a given value of S and room-temperature current, the ambient temperature at which this departure first becomes significant depends on the magnitude of V_B.

Moreover, because of the avalanche factor, M, the idea of reverse turnover voltage being doubled if the coefficient of thermal drop, S, could be halved[6] is erroneous. (For example, at room temperature, the theoretical V_1 approaches V_B just the same.) Nevertheless it is useful to reduce S as much as possible, since the effect of a low value of S on increasing V_1 becomes more pronounced at higher ambient temperatures, and, when surface leakage is also considered (see Section 6), may be important even at room temperature.

(3.4.4) **Approximate Form of I_s.**

Expression (29) can be conveniently approximated by

$$I_s = I_0 \exp\left(\frac{T - T_0}{\phi}\right) \quad . \quad . \quad . \quad . \quad (37)$$

where I_0 is the current at $T = T_0$, and ϕ is a constant. [$\phi \simeq 12{\cdot}5°\,\mathrm{C}$ for germanium alloyed junctions, and this approximation is then valid as compared with eqn. (29) over a range of about 100° above $T_0 = 300°\,\mathrm{K}$ at the 5% level of error of calculation of T from given I_s values.]

The corresponding turnover analysis yields $\theta_1 = \phi = \text{constant}$, as compared with the true value of θ_1 specified by eqn. (34), and $I_1 = M_1I_s(T_a)\varepsilon$, where ε is the base of the natural logarithms.

(4) VARIATION OF TURNOVER POINT WITH TEMPERATURE

A graphical interpretation of the solutions obtained above is presented in Figs. 3 and 4. It should be remembered that the diagrams are valid only for the case where the saturation current consists entirely of a bulk component. To simplify matters, the approximation in Section 3.4.4 is used throughout, with $n = 3$.

The expression (36) implicitly contains the four parameters S, ϕ, V_B and $I_s(300°\,\mathrm{K})$ which are peculiar to any particular diode. To make the curves universally applicable, the constant $A = \phi/[\epsilon V_B S I_s(300°\,\mathrm{K})]$ must first be calculated for the particular diode under consideration. Then eqn. (36) becomes

$$\frac{V_1}{V_B}\exp\left(\frac{T_a - 300}{\phi}\right) = A\left[1 - \left(\frac{V_1}{V_B}\right)^3\right] \quad . \quad . \quad (38)$$

In Fig. 3 the full line is a plot of A against ambient temperature in units of ϕ, corresponding to $V_1 = 95\%V_B$, showing when the distinction between breakdown and turnover first becomes noticeable theoretically; e.g. for typical values of $\phi = 12{\cdot}5°\,\mathrm{C}$, $S = 0{\cdot}35\,\mathrm{deg\,C/mW}$, $I_s(300°\,\mathrm{K}) = 4\,\mu\mathrm{A}$, and $V_B = 100$ volts (i.e. $A = 33$), this would occur at $T_a = 320°\,\mathrm{K}$, whereas for 5% departure of V_1 from V_B to occur at 300° K ambient, V_B would have to be about 500 volts. At the other extreme, M_1V_1 is governed principally by V_1 to 5% accuracy when $M_1 \leqslant 1{\cdot}05$ (i.e. whenever $V_1 \leqslant 0{\cdot}365V_B$ for $n = 3$). The broken line in Fig. 3 shows the variation of T_a with A at the edge of this pre-breakdown region ($V_1 = 36{\cdot}5\%\ V_B$).

In Fig. 4, for a given value of A (33, using above figures), it is shown how V_1/V_B decreases as the ambient temperature, again plotted in units of ϕ, is raised. For a specified value of V_1/V_B and any other value of A, this diagram can be used to calculate T_a by adding to the value of $(T_a - 300)/\phi$ obtained from the graph the amount $\log_e(A/33)$. Fig. 4 can also be used to design for A. A typical example will make this clear.

Suppose the crystal material of the diode is such that $\phi = 10°\,\mathrm{C}$, $V_B = 125$ volts and $I_s(300°\,\mathrm{K}) = 4\,\mu\mathrm{A}$, and it is decided that

* When the first approximation (24) is considered, the formula (34) for θ_1 is still valid provided that b' is replaced by $b'' = (W_G - W_A)/2k$. The magnitude of θ_1 when the intrinsic range is reached is thus seen to be at least double its extrinsic value. With the figures just considered, except for $m = 0{\cdot}85$, $(\theta_1)_{int} = 44°\,\mathrm{C}$.

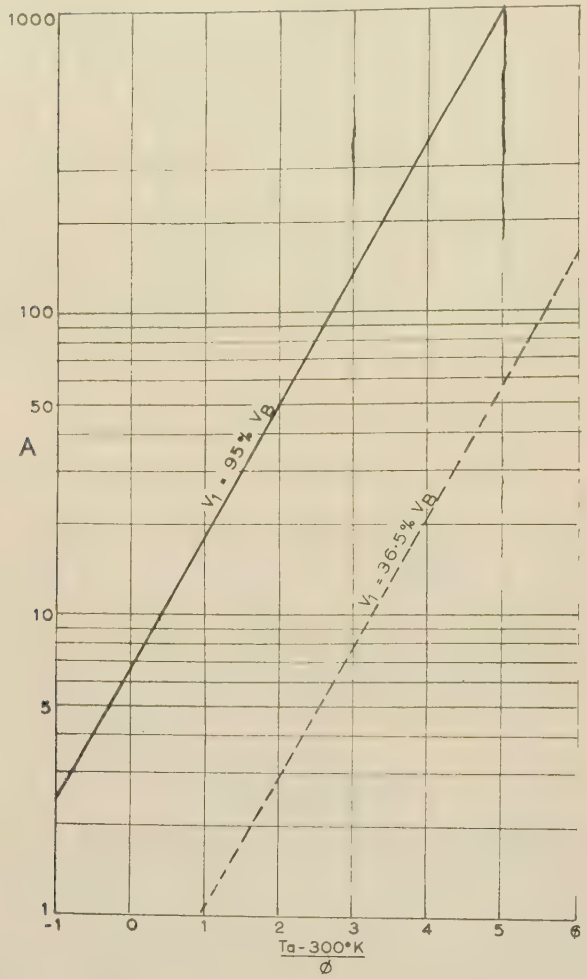


Fig. 3.—Variation of ambient temperature with device constant A.
—— 5% deviation of V_1 from V_B.
----- Edge of pre-breakdown region.

a junction temperature $T_{jmax} = 360°\,\mathrm{K}$ must not be exceeded. It is required to specify the thermal resistance necessary in order that the operating voltage may be 50 volts, say, when a constant voltage supply is employed, and a 'theoretical' (i.e. no surface leakage) junction is assumed.

With $T_1 = 360°\,\mathrm{K}$, the maximum permissible ambient temperature is $T_{amax} = T_1 - \phi = 350°\,\mathrm{K}$, so that $(T_{amax} - 300)/\phi = 5$. With $V_1 = 40\%V_B$, Fig. 4 gives a value of $(T_a - 300)/\phi = 4{\cdot}35$. Hence the required value of A is given by $\log_e (A/33) = 0{\cdot}65$, i.e. $A \simeq 63$. This yields a maximum value of S of approximately $0{\cdot}12\,\mathrm{deg\,C/mW}$.

The curvature at the turnover points can be shown to be

$$\left(\frac{d^2V}{dI^2}\right)_1 = \frac{-S^2V_1^2\left(\frac{\partial^2 I}{\partial T^2}\right)_1}{\left(\frac{\partial I}{\partial V}\right)_1 + SI_1\left(\frac{\partial I}{\partial T}\right)_1} \quad . \quad . \quad . \quad (39)$$

and its value will depend on $I_s(T_a)$, the initial ambient-temperature current, and S, the thermal resistance.

(5) PARAMETERS REQUIRED TO CALCULATE THE STEADY-STATE CURVES

To establish a link between theory and experiment, the turnover characteristics calculated from a knowledge of V_B, n, S, $I_s(T_a)$ and ϕ for a given junction can be compared directly with a set of d.c. readings, using a high-impedance source. Care must be taken not to exceed a specified junction temperature, T_{jmax}, which will depend on the metallurgical and mechanical structure of the unit.

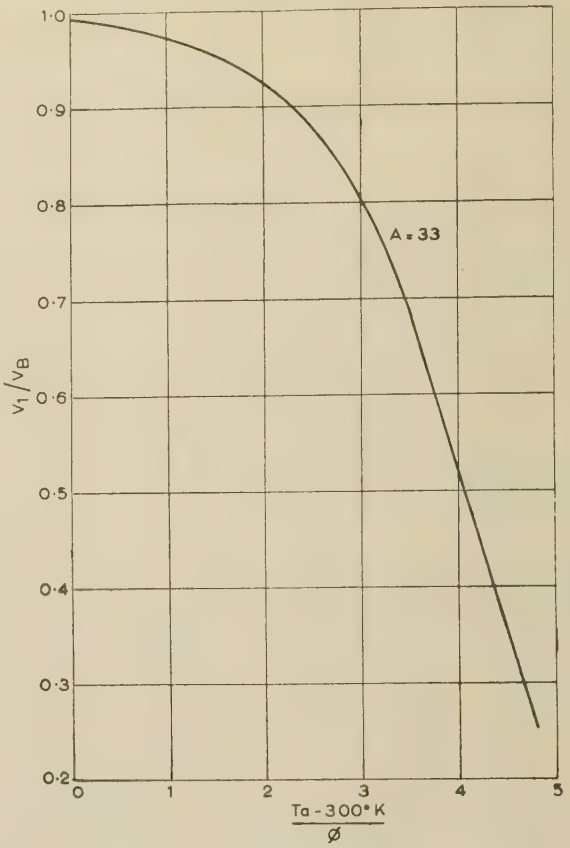


Fig. 4.—Variation of ambient temperature with turnover voltage for a fixed value of A.

(5.1) Theory and Experiment of V_B

In the past, empirical formulae of the type

$$V_B = a\rho_n + b\rho_p \quad . \quad . \quad . \quad . \quad . \quad (40)$$

for step junctions have been based on the theory of spontaneous Zener emission[20] when a critical field has been reached within the depletion layer. More recently, however, Miller has shown[5] that breakdown is governed by an avalanche multiplication process similar to Townsend's β-mechanism in gases,[21] with different effective ionization rates for holes and electrons, leading to a non-linear relationship of the form

$$V_B = K\rho^{0{\cdot}725} \quad . \quad . \quad . \quad . \quad . \quad (41)$$

where ρ is the resistivity of the high-resistivity side of junction and it is assumed that the resistivity of the other side is negligible in comparison with it, and K is a constant. Alternatively, V_B can be determined by extrapolation of the data relating the ionization rates to the maximum field in the barrier region, published by McKay and McAfee.[22] Knott *et al.*[23] found the avalanche process to be applicable to n-type base material of $\rho_n > 0{\cdot}5$ ohm-cm, but below this the Zener field-emission theory was applicable.

To determine V_B experimentally, measurements should be made by pulsing methods, in which the length of the pulse should

exceed the electric time-constant of the junction, yet be less than the thermal one.

(5.2) Theory and Experiment of S, I_s and φ

S depends on cooling arrangements, and values as low as 0·005 deg C/mW have been reported[17] for certain power transistors.

I_s is controllable by doping the base to the required impurity concentration, and also by choice of junction area.

ϕ lies in the range $10 < \phi < 16$ for germanium alloyed junctions, but its value may change slightly when the approximation (37) ceases to be valid over the complete temperature range investigated. Roughly speaking, the lower limit corresponds to a diode with $W_A \simeq 0$, while the upper limit corresponds to the opposite extreme of a transistor junction with high life-time activation energy ($W_A \simeq 0{\cdot}3$ eV); however, in the latter case, the proximity of the floating junction is significant, and values of ϕ have been found to differ between the collector and emitter junctions of the same transistor.

(5.3) Use of p–n–p Transistors to Determine n

If α_{N0} denotes the low-voltage current gain in the common-base configuration, the common-emitter breakdown voltage is related[5] to V_B by

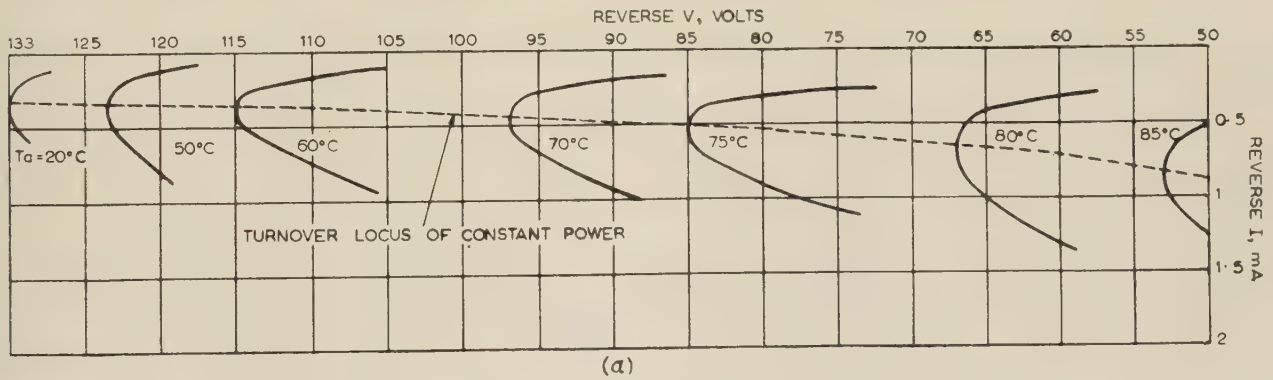
$$\left(\frac{V_k}{V_B}\right)^n = 1 - \alpha_{N0} \quad . \quad . \quad . \quad . \quad . \quad (42)$$

This expression yields the best way of determining n. With the base floating, measurements of I', the emitter-to-collector through current, can be made for different reverse voltages, V'.

Although α_{N0} may be strongly dependent on $I_E \simeq I'$ when the base is floating, this does not affect the linearity of log plots of V_k/V_B against $1 - \alpha_{N0}$, provided that the same chosen value of I_E is applied to each transistor,[24] giving a line of slope n when several transistors of the same resistivity type are taken. Whether this latter line passes through the point ($\alpha_{N0} = 0$, $V_k/V_B = 1$) is at the same time an indication of whether the breakdown voltages observed are due to bulk or to surface effects. For further information the reader is referred to the paper by Miller,[5] which describes this method in more detail.

(6) MODIFICATION OF ANALYSIS TO INCLUDE SURFACE LEAKAGE CURRENT

Fig. 5(*a*) shows a set of theoretical steady-state curves, calculated according to the above analysis neglecting surface leakage, and Fig. 5(*b*) shows typical corresponding experimental curves on the same scale, relating to the collector junction of a p–n–p transistor. It is seen that the experimental curves compare favourably with the theoretical ones only at the higher ambient temperatures, while there appears to be a marked discrepancy at the lower temperatures (e.g. room temperature). This is almost certainly due to surface leakage current, which has received much attention in the literature,[25] and no attempt is made here to analyse it accurately. It should be pointed out, however, that a finite time—of the order of seconds to minutes—is required for surface states to come to equilibrium with the bulk, so that calculation of leakage current, I_L, is strictly valid only for the d.c. case.



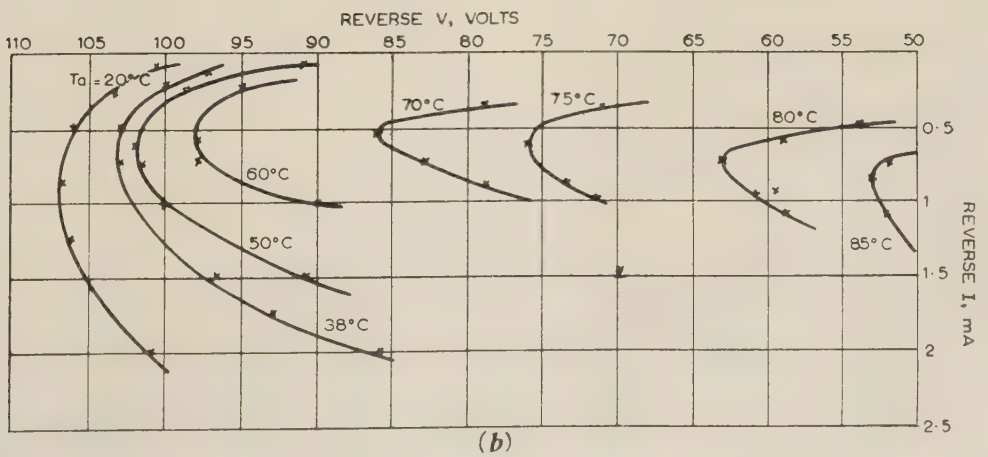


Fig. 5.—Steady-state reverse characteristics with $S = 0{\cdot}35$ deg C/mW, $\phi = 15°$ C, $n = 3$, $I_s(20°\,\text{C}) = 3{\cdot}5\mu$A and $V_B = 133$ volts.
(*a*) Theoretical. (*b*) Experimental.

Let the diode reverse current now be written as

$$I = MI_s + I_L \quad . \quad . \quad . \quad . \quad . \quad . \quad (43)$$

where

$$I_L = I_L(V) \exp\left[\frac{\lambda(T - T_a)}{\phi}\right] \quad . \quad . \quad . \quad . \quad (44)$$

and $0 \leqslant \lambda < 1$, so that $I_L(V)$ is the voltage-dependent part of I_L, and the extreme values of λ (0 and 1) would refer respectively to

(*a*) I_L being independent of temperature.
(*b*) I_L having the same temperature variation as I_s.

Although case (*b*) has not been observed in practice,[26] $\lambda = 1$ might be useful as a safe upper limit in designing for the turnover point. The form given in eqn. (44) with a value of $\lambda = 0{\cdot}5$ has been derived for a very simple model of surface leakage current by Cutler and Bath.[27] The flexibility of form (44) of I_L with respect to T is introduced here because the relationship[27] is 'usually quite variable, even between units of a given type'.

Substitution from eqns. (43) and (44) into eqn. (16) shows that the equation $\theta_1 = \phi$ must be replaced by

$$\theta_1 = T_1 - T_a = \phi\left(\frac{M_1 I_{s1} + I_{L1}}{M_1 I_{s1} + \lambda I_{L1}}\right) \quad . \quad . \quad . \quad (45)$$

where $I_{L1} = I_L(V_1) \exp \lambda\left(\frac{T_1 - T_a}{\phi}\right)$

and eqn. (36) is replaced by

$$V_1 = \frac{\phi}{S(M_1 I_{s1} + \lambda I_{L1})} \quad . \quad . \quad . \quad . \quad (46)$$

It is interesting to consider the extreme possible values of λ (0 and 1) in the light of eqns. (45) and (46). Were $\lambda = 0$, the highest value of θ_1 would be obtained at the lowest (e.g. room) temperature, since I_{L1}/MI_{s1} might be appreciable, but θ_1 would decrease towards ϕ as the ambient temperature was raised. However, V_1 would then be no different from the theoretical value corresponding to the case of no surface leakage, previously discussed. With $\lambda = 1$, θ_1 would equal ϕ at any temperature, but V_1 could then be much lower than the predicted value, especially at the lower ambient temperatures.

The fact that the latter behaviour of V_1 is observed, coupled with the approach of the expected values of V_1 at the higher ambient temperatures [Fig. 5(*b*)], suggests that, whilst λ must not be zero, the temperature variation of I_s is much greater than that of I_L, so that λ cannot equal unity either. An intermediate figure like $0{\cdot}5$ is therefore more in keeping with the observed turnover characteristics.

Eqn. (46) also shows that, with a known value of ϕ, turnover powers can no longer be used to measure S directly, but surface leakage power dissipated must first be subtracted. Presumably, the experimental verification[3] of the constancy of θ_1 ($\simeq 11°$ C) at turnover (or what amounts to the same thing, runaway with a constant-voltage source) was performed on units which exhibited no surface leakage effects.

(7) EXPERIMENTAL PROCEDURE

To obtain ambient temperatures reasonably constant, the encapsulated junctions investigated were enclosed in a standard laboratory oven, which controlled temperature within $\pm 1°$ C. Isothermal characteristics for determining V_B were obtained at different elevated temperatures with the aid of a pulse generator[29] built in the laboratory. Pulses of 75 microsec duration and controlled amplitude were applied across the series combination of the diode and a 2·5-kilohm resistor. The potential difference across the combination and that across the series resistor were measured by an oscillograph.

The steady-state readings were obtained by means of a pentode high-impedance d.c. source and a valve voltmeter having negligible shunting effect. It was found that several minutes were required at each point of the characteristic before a definite reading could be obtained, but the static characteristics could be retraced with very little hysteresis. The long time-constant may have been due to thermal effects, or it may have been associated with the finite time required for the surface to reach equilibrium with the bulk, as mentioned previously. Thus, although negative-resistance portions of the characteristic could be reached, applications involving a swing about such an operating point would be possible only at extremely low frequencies.

(8) ANALYSIS OF EXPERIMENTAL RESULTS

The characteristics plotted in Fig. 5(*b*) relate to the collector junction of a *p–n–p* transistor, where ϕ was found by measuring the low-voltage variation of I_s with T. The base material was of such resistivity that only a small departure from the extrinsic range (and therefore deviation of ϕ from 15°) was observed, even at the highest ambient temperatures. S was determined by the use of the conventional 'time division' technique described by Ollendorf and Loofbourrow.[28]

The value of λ was then calculated as follows:

The observed value of V_1, which was clearly defined and should be accurate to 5%, was substituted into eqn. (46) and from a knowledge of S and ϕ the value of $M_1 I_{s1} + \lambda I_{L1}$ could be calculated ($= I'_{L1}$, say). This was, in turn, substituted into eqn. (45), together with the measured overall value of I_1 ($= M_1 I_{s1} + I_{L1}$), to yield a value of T_1. (Unfortunately, owing to the small curvature of the characteristic at the turnover points of the junctions investigated, the depicted value of I_1 could be in error by as much as 15%. An average value of I_1, however, for each curve, gave λ more or less constant, as Table 1 shows.)

Table 1

ANALYSIS OF TURNOVER POINTS

T_a (measured)	V_1 (measured)	I_1 (measured)	M_1	I_{L1}	$M_1 I_{s1}$	$T_1 - T_a$	T_1	$I'_{L1} = \phi/SV_1$	$I_L(V_1)$	λ
deg C	volts	mA		mA	mA	deg C	deg C	mA	mA	
20	107	0·9	2·11	0·83	0·07	34	54	0·40	0·31	0·40
38	103	0·75	1·87	0·62	0·13	27	65	0·42	0·28	0·465
50	102	0·65	1·83	0·43	0·22	23	73	0·42	0·21	0·47
60	98	0·6	1·63	0·28	0·32	20·5	80·5	0·44	0·15	0·43
70	86	{0·6 0·55	1·40	{0·16 0·165	{0·44 0·385	{17·5 16·5	{87·5 86·5	0·50	0·09(5)	{0·38 0·7
75	76	0·65	1·23	0·115	0·535	17	92	0·565	0·07	?
80	63	0·75	1·16	0·08	0·67	16·5	96·5	0·68	0·05	?
85	53	0·85	1·06	0·04	0·81	16	101	0·81	0	?

In the tenth column $I_L(V_1)$ is the voltage-dependent part of I_L, calculated with λ taken as 0·45.

From T_1, I_{s1} was calculated on the basis of eqn. (37), whilst M_1 could be written down from a knowledge of V_1 and V_B. Then the equations

$$I_{L1} = I_1 - M_1 I_{s1}$$

and

$$\lambda = \frac{I'_{L1} - M_1 I_{s1}}{I_{L1}}$$

were used to determine I_{L1} and λ in turn. This method is less accurate for the higher ambient temperatures, where the calculated value of λ is more sensitive to errors in the measurement of I_1. The average value of λ of 0·45 computed from the measurements is very closely in accord with the theory given by Cutler and Bath,[27] who verified the theoretical figure of $\lambda = 0{\cdot}5$ experimentally for silicon diodes.

(9) CONCLUSIONS

From Table 1 it can be seen that the value of θ_1 does approach the theoretical limit of ϕ (= 15° C) at the highest ambient temperature (85° C), where the position of the turnover point is almost identical with the theoretical one, i.e. without surface leakage [see Fig. 5(*a*)], while the largest difference (room temperature of 20° C) is the result of θ_1 decreasing monotonically with T_a, as suggested by eqn. (45). Also, the increase of $I_L(V_1)$ with V_1 is in agreement with the theory on this subject (e.g. simple inversion layers would give logarithmic variation[30] with V, whereas the effect of water vapour at the surface would tend to produce a linear variation[31]), and observations of the multiplicative nature of $I_L(V)$ at the higher voltages.[32] The constancy of the values of λ calculated from the measurements is consistent with the assumption of form (44), while the decrease in V_1 with T_a, slowly at first but more rapidly as the ambient temperature is increased, is typical and to be expected from the theory.

The usefulness of this semi-quantitative analysis lies in the derivation of conditions governing the avoidance of runaway, thus permitting the determination of a maximum working voltage and also a minimum series resistance suitable for diode operation. The necessary information is furnished by the static curves and is more easily interpreted when surface leakage is small.

At the same time, the analysis should serve as an introduction to the conditions applicable to transistor operating points. From the diode theory, and from the figures presented in Table 1, turnover typically occurs before the maximum permissible junction temperature is reached. Whereas, in the case of the diode, the $T_{j\,max}$ limitation referred to a single point only on the characteristic, since no degrees of freedom remain in eqns. (14) and (15) when T is specified, in the case of a transistor, a similar analysis might restrict the region of operation to an area well within the hyperbola $T_{j\,max} - T_a = SI_cV_c$ of constant power.

(10) ACKNOWLEDGMENTS

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(12) APPENDICES

The runaway condition given in eqn. (20) relating to the switch-on of a reverse-biased diode is derived here, together with the equivalence of eqns. (19) and (20) for the intersection points. The analysis is restricted to the simple case of a diode in series with a battery and a limiting resistor.

(12.1) Derivation of Condition (20)

The equation of the load line is

$$V = V_{cc} - IR \quad . \quad . \quad . \quad . \quad . \quad . \quad . \quad (47)$$

The reverse current can be written as in eqn. (37), namely

$$I = MI_s(T_a)\exp\left(\frac{T - T_a}{\phi}\right) \quad . \quad . \quad . \quad . \quad (48)$$

If steady-state conditions can be reached, then after a sufficient time has elapsed for thermal equilibrium to be attained, the additional eqn. (8), rewritten here for completeness also holds:

$$T - T_a = SP \quad . \quad . \quad . \quad . \quad . \quad . \quad (8)$$

where $P = IV$.

In this case, eqns. (47), (48) and (8) taken together refer to a single point (i.e. a stable intersection of the load line with the static characteristic).

To investigate whether thermal stability will prevail when the diode is first switched to any other point along a load line, eqn. (8) is replaced by the stability criterion

$$dT/dP > S \quad . \quad . \quad . \quad . \quad . \quad . \quad (49)$$

Eqns. (47) and (48) can then be used to calculate dT/dP as follows:

Taking total differentials of eqn. (48) gives

$$dI = I_s(T_a)\exp\left(\frac{T - T_a}{\phi}\right)\left(\frac{\partial M}{\partial V}dV + \frac{M}{\phi}dT\right) \quad . \quad (50)$$

neglecting $\partial M/\partial T$ (see Section 3.4.2).

Substituting $dV = -RdI$ from eqn. (47) into eqn. (50) and solving for dT yields

$$dT = \frac{\phi}{I}\left(1 + \frac{IR}{M}\frac{\partial M}{\partial V}\right)dI \quad . \quad . \quad . \quad . \quad (51)$$

Multiplying both sides of eqn. (47) by I and differentiating gives

$$dP = (V_{cc} - 2IR)dI \quad . \quad . \quad . \quad . \quad (52)$$

Eliminating dI between eqns. (51) and (52) gives

$$\frac{dT}{dP} = \frac{\phi}{I}\frac{\left(1 + \frac{IR}{M}\frac{\partial M}{\partial V}\right)}{V_{cc} - 2IR} = \frac{\phi}{I}\frac{\left(1 + \frac{IR}{M}\frac{\partial M}{\partial V}\right)}{V - IR} \quad . \quad (53)$$

and

$$\frac{1}{M}\frac{\partial M}{\partial V} = \frac{n}{V}(M - 1) \quad . \quad . \quad . \quad . \quad . \quad (54)$$

by differentiation of eqn. (4).

On replacing $\frac{1}{M}\frac{\partial V}{\partial M}$ in eqn. (53) by the value in eqn. (54) and substituting for dT/dP into eqn. (49), an upper limit is obtained to the power dissipated in the diode in excess of that dissipated in the resistance, expressed by

$$IV - I^2R < \frac{\phi}{S}\left[1 + \frac{IRn}{V}(M - 1)\right] = K, \text{ say} \quad . \quad (20)$$

The function K will be roughly constant for a constant-voltage source, or whenever M is sufficiently close to unity. Inequality (20) divides the (I, V) plane into two regions, stable and unstable, for a given load, R, and thermal resistance, S [e.g. for the constant-voltage source, $R \simeq 0$, so that the condition becomes $IV < \phi/S$, the boundary thus being a rectangular hyperbola, as shown in Fig. 2(*b*)].

(12.2) Equivalence of Eqns. (19) and (20) at the Points of Intersection of the Load Line with the Steady-State Curve

The slope at a point (V, I) along the steady-state curve is obtained by substituting for $T - T_a$ from eqn. (8) into eqn. (48) [eqn. (47) no longer holds] to give

$$I = MI_s(T_a)\exp\left(\frac{SIV}{\phi}\right) \quad . \quad . \quad . \quad . \quad (55)$$

Differentiation with respect to V, taking into account the dependence of I upon V in the exponent, and solving for dV/dI yields

$$r = \frac{dV}{dI} = \frac{1 - \frac{S}{\phi}IV}{\frac{I}{M}\frac{\partial M}{\partial V} + \frac{S}{\phi}I^2} \quad . \quad . \quad . \quad . \quad (56)$$

Solving eqn. (56) as a linear equation in S gives

$$S = \frac{\phi\left(1 - r\frac{I}{M}\frac{\partial M}{\partial V}\right)}{I(V + Ir)} \quad . \quad . \quad . \quad . \quad (57)$$

Substituting into eqn. (49) this value of S and the value of dT/dP from eqn. (53), from which eqn. (20) was derived, gives

$$(r + R)\left(1 + \frac{V}{M}\frac{\partial M}{\partial V}\right) > 0 \quad . \quad . \quad . \quad . \quad (58)$$

on simplification.

Since the sum in the second bracket is essentially positive, the stability condition for points on the static curve is

$$r + R > 0 \quad . \quad . \quad . \quad . \quad . \quad . \quad (19)$$

Hence eqns. (19) and (20) are equivalent for such points [e.g. B, D or F of Fig. 2(*a*)].

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May 1957

TRANSISTOR CIRCUITS AND APPLICATIONS

By A. G. MILNES, D.Sc., Associate Member.

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SUMMARY

Junction-transistor circuits are reviewed, and new concepts and trends in design are discussed. Improvements in transistor ratings are described, but special transistor types such as tetrodes, *p-n-p-n* and unipolar devices are not considered.

Basic applications surveyed include low-level and power amplifiers, square-wave and sinusoidal oscillators, power convertors and regulators, waveform-shaping and computing circuits, and low-level and power-switching circuits. Numerous references are given.

Transistors are capable of supplanting thermionic circuits in all except high-frequency applications (above 10 Mc/s with currently-available units). High temperatures will continue to present limitations in military applications, until silicon transistors become widely available. In more general applications, such as the entertainments field, cost is at present an important limiting consideration.

LIST OF PRINCIPAL SYMBOLS

f = Frequency, c/s.
f_α = Common-base α cut-off frequency, c/s.
I_c = Collector current, amp.
I_{c0} = Collector leakage current, amp.
K = Ratio of signal applied to signal amplitude for full output.
N = Number of winding turns.
P_d = Circuit power loss to be dissipated from transistors, watts.
P_l = Alternating power output in the load, watts.
t = Time, sec.
V_b = Bias-voltage amplitude, volts (peak).
V_c = Collector-to-emitter voltage rating, volts.
V_s = Signal amplitude, volts (peak).
α = Current gain (collector-to-emitter).
θ = Switching angle, rad.
ϕ = Instantaneous core flux, maxwells.
Φ_s = Saturation core flux, maxwells.
ω = α cut-off angular frequency, rad/sec.

(1) INTRODUCTION

Transistor circuit technology has developed considerably in the last two or three years, and these developments and circuit trends are reviewed. A basic knowledge of transistor circuit principles and characteristics is presumed, since textbooks are available presenting the fundamentals of the subject.

One major factor in the progress recently achieved has been the general availability of transistors of acceptable performance characteristics and reliability. In particular, the point-contact type of construction has been superseded by alloyed- or grown-junction units, and the circuits discussed are confined solely to junction-transistor triodes. Special transistors such as tetrodes, *p-n-p-n* and unipolar units are not considered in the paper since these devices are available only to a limited extent.[1]

Improvements in junction-triode characteristics have been substantial as manufacturing control and processing techniques have developed. Collector-voltage ratings in germanium units have risen from 5 to 30 volts, and constructions originally limited to a few milliamperes have had the current ratings increased by about an order of magnitude as improvements in emitter injection efficiency have been made. Ring-base constructions have become fairly general, and they give reduced base resistance, improved gain and better heat-transfer characteristics. Alloy units with α cut-off frequencies in the range 5–10 Mc/s and surface-barrier types in the range 50–100 Mc/s are now available, and higher-frequency units are under development.

Power transistors have become available in germanium with current ratings of several amperes and power-handling capacities of tens of watts. The overall thermal gradients of such units on good heat sinks may be as low as 5° C per watt, and so power dissipations of many watts become permissible. Maximum junction temperatures of about 75° C are quoted for germanium transistors normally available at present; if such units are held at higher junction temperatures some deterioration of characteristic occurs in the course of a few hundred hours. Such deterioration appears to be related to surface conditions, and since these are coming more and more under control, manufacturers are beginning to list types with permissible junction temperatures of 90–100° C. However, it must be stressed that for fundamental physical reasons transistor parameters, such as I_{c0} and α, vary considerably with junction temperature, and such changes present problems in circuit design that may limit the temperature range over which particular amplifiers, oscillators, etc., function satisfactorily.

From energy-gap considerations, silicon is basically more suitable for high-temperature operation than germanium, and silicon transistors for 150° C junction temperature operation are available. Again, the variations of parameters with temperature present limitations in circuit design, e.g. in silicon power transistors there may be loss of gain at low temperatures such as −50° C.

Information on transistor reliability essentially requires large-scale long-term tests in practical equipment, with samples taken from production at frequent intervals to give statistical validity. The limited information available from digital-computer and telephone systems shows failure rates of between 0·1 and 1% per 1 000 hours. Such figures are comparable with the results for premium-grade thermionic valves developed for high reliability, and must be regarded as very encouraging.

(2) AMPLIFIERS

(2.1) Power and Dissipation Ratings

Transistor power amplifiers for the operation of small motors and relays in servo-type applications are of considerable potential interest with the recent availability of transistors of higher power ratings. However, many such applications are subject to ambient temperatures as high as 55° C, and since the maximum recommended junction temperature for germanium transistors may be 75° C, circuits of high efficiency must be used so that

This is an 'integrating' paper. Members are invited to submit papers in this category, giving the full perspective of the developments leading to the present practice in a particular part of one of the branches of electrical science.

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the power loss and temperature rise in the transistor are low relative to the power handled.

Class-A circuit configurations of common base, common emitter and common collector differ in respect of input and output impedances and gain as described in standard textbooks.[2, 3] If the rated collector-emitter voltage is V_c and the peak collector current is I_c, the quiescent class-A operating point is normally set at the co-ordinates $\frac{1}{2}V_c$, $\frac{1}{2}I_c$. For full a.c. signal applied, the alternating output power obtained is approximately $\frac{1}{8}V_cI_c$,

namely
$$P_l = \left(\frac{1}{\sqrt{2}}\frac{V_c}{2}\right)\left(\frac{1}{\sqrt{2}}\frac{I_c}{2}\right). \quad . \quad . \quad . \quad . \quad . \quad (1)$$

By simple integration the transistor dissipation may be shown to be equal to the load power, and the efficiency is therefore 50%. In the quiescent condition (i.e. without signal applied) the transistor dissipation is, of course, determined by the co-ordinates of the quiescent point and is therefore $\frac{1}{4}V_cI_c$.

Class-B operating conditions have been considered in detail by Hillbourne and Jones.[5] Maximum gain is given by the common-emitter arrangement, which in simple form is shown in Fig. 1(*a*). The dot symbols indicate the winding sense, where

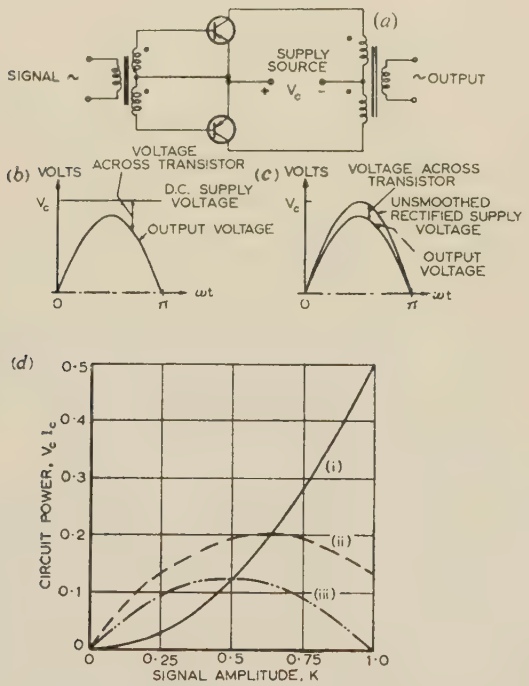


Fig. 1.—Class-B transistor output-stage performance.
(*a*) Common-emitter class-B circuit.
(*b*) Operation with d.c. supply source.
(*c*) Operation with unsmoothed rectified supply.
(*d*) Theoretical power-output and power-loss curves.
(i) Power output.
(ii) Power loss (battery source).
(iii) Power loss for rectified unsmoothed a.c. source.

this is relevant. With a battery supply source, the expressions for a.c. load power output and for the dissipation loss of the two transistors together are, respectively,

$$P_l = K^2\frac{V_cI_c}{2} \quad . \quad . \quad . \quad . \quad . \quad . \quad (2)$$

and
$$P_d = \frac{K}{\pi}\left(2 - \frac{K\pi}{2}\right)V_cI_c \quad . \quad . \quad . \quad . \quad . \quad (3)$$

where K is the fraction of full signal considered to be applied. During its non-active half-cycle each transistor must be capable of withstanding a voltage $2V_c$, equal to twice the applied battery voltage, because of the class-B doubling action of the output transformer.

Curves derived from eqns. (2) and (3) are given in Fig. 1(*d*). The maximum power output into the load is $\frac{1}{2}V_cI_c$ at an efficiency of 78%, which may be compared with $\frac{1}{4}V_cI_c$, at 50% efficiency, obtainable from two such transistors connected in parallel and operated in class-A conditions. In considering the required dissipation rating of the class-B arrangement for a given power output rating it is important to note that the power dissipation reaches a maximum ($0{\cdot}202V_cI_c$) for an output condition less than the full swing (namely at $K = 0{\cdot}64$).

It has been recognized recently that in class-B connections there are advantages to be gained in operation from an unsmoothed rectified sinusoidal supply source, rather than a direct-voltage supply such as a battery.[6] The rectified supply should be derived from a sinusoidal wave of the same frequency and phase as the signal (this is quite practical in synchro-fed servo systems), and the peak amplitude of the supply should be equal to the rated voltage, V_c, of the transistor. This form of supply reduces the voltage sustained across the transistor between collector and emitter [cf. Figs. 1(*b*) and 1(*c*)] and so reduces the power loss in the transistor. In fact, for the full signal condition the source voltage provides exactly the voltage waveform required for full output and there is no power loss in the transistor (except for the base drive dissipation, which is small). For a fraction, K, of full signal, the load- and dissipation-power expressions are as follows:

$$P_l = K^2\frac{V_cI_c}{2} \quad . \quad . \quad . \quad . \quad . \quad . \quad (4)$$

as for normal class-B operation, and

$$P_d = \frac{K}{2}(1 - K)V_cI_c \quad . \quad . \quad . \quad . \quad . \quad (5)$$

The maximum dissipation loss is $0{\cdot}125V_cI_c$, which occurs at $K = 0{\cdot}5$, and this is appreciably less than the loss of $0{\cdot}202V_cI_c$, which occurs with normal d.c. supply conditions.

(2.2) Improvements in General Circuit Technique

Aspects of transistor-amplifier design that have been the subject of recent study include neutralization, local and overall feedback shaping techniques, automatic gain control, correction of zero drift by monitoring and techniques for achieving stability under high-temperature conditions.

Neutralization techniques aim at converting transistor circuits from bilateral to unilateral, so that internal feedback effects between output and input are reduced.[7] Typically this may require the addition of a capacitor and resistor for each transistor stage. Neutralization is a useful aid in amplifier design, since the problems of matching complex load and signal-source impedances through a bilateral device are difficult and at the very least involve cumbersome analysis. On the other hand, neutralization techniques require additional circuit components that may be regarded as not indispensable.

Since transistors may be used in various possible connections (common base, common emitter and common collector) and there are several techniques for coupling between transistor stages in a typical three- or four-stage amplifier, comparative circuit studies are valuable. De Sautel's examination of the relative merits of transformer coupling, resistance-capacitance coupling and direct coupling between stages of transistors in common-emitter connection is an example of the type of circuit assessment involved.[8, 9] Direct coupling at low collector

voltages has been studied by Beter *et al.*,[10] and shown to give useful circuits for computer purposes.

Local and overall feedback techniques to achieve the required gain and phase characteristics for given signal frequency ranges have been described with reference to operational amplifiers for analogue computers.[11] In transistor amplifiers the magnitude of the overall negative feedback possible without instability tends to be limited by the frequency characteristics of the transistor. However, with suitable phase-advance shaping in the feedback path, Almond and Boothroyd[12] have shown, for example, that amplifier gains of 33 dB with 30 dB negative feedback are possible.

Automatic gain control of transistor amplifiers is possible by the provision of feedback from the output to an input-stage transistor in such a way that increasing output decreases the emitter current or the collector voltage of the controlled transistor, and so decreases the gain. The principles involved have been reviewed in detail by Chow and Stern, who conclude that automatic gain control involves problems of distortion, bandwidth and centre-frequency variation, but that in many applications these can be solved by appropriate design.[13]

Zero drift is usually appreciable in d.c. transistor amplifiers, and in precision amplifiers it must be removed by zero-set techniques such as that proposed by Goldberg for d.c. thermionic amplifiers.[14] Basically the method is a comparison between the input and output quantities to reveal any zero drift or gain change as a d.c. difference signal which is then converted to alternating current by a mechanical 'chopper' or a magnetic modulator and fed into a stable low-drift a.c. amplifier. The amplified a.c. output is reconverted to direct current and applied to the input of the controlled amplifier as a drift-correcting signal. Blecher has applied this procedure to transistor amplifiers in summing and integrating applications,[11] and for the temperature range 0–50° C, he has reduced drifts of $\pm 1 \cdot 5$ volts output to less than ± 5 mV.

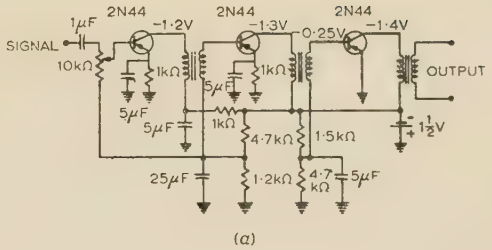
The behaviour of transistor circuits over wide temperature ranges is of importance to many potential users. Germanium transistor amplifiers stable to 95° C have been described by Greatbatch and Hirtreiter, who use supplementary transistors to stabilize the collector-emitter voltages of the main transistors.[15] When applied to silicon transistors the technique is reported to give stability up to 135° C. However, the problem is not completely solved merely by devising circuits that do not burn-out at the highest temperatures, since problems of change of circuit gain and balance over a wide temperature range remain.[16, 17, 18]

(2.3) Amplifier Applications

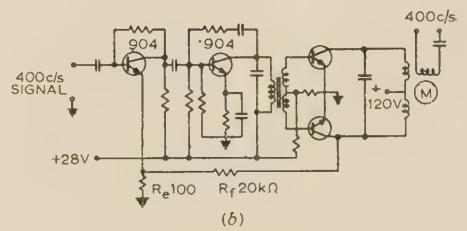
Applications of transistor amplifiers described in recent papers cover hearing aids, radio receivers, radar video circuits, and instrument and servo-mechanism circuits. Although a detailed review of the circuits would be out of place, a few performance figures will be given as an indication of what can reasonably be achieved.

The simple one-battery hearing-aid circuit in Fig. 2(*a*) has a gain of 76 dB and provides an output of 4·8 mW for a total battery input of 16·5 mW; the overall efficiency is thus 29%. The 3 dB points on the frequency-response curve give a bandwidth of 200–5 000 c/s. Some improvement is possible if a two-battery circuit is used.[19]

Radio receivers using transistors are commercially available, and typical circuits are described in the literature.[20–29] For a pocket-size receiver the power dissipation is 20 mA at 5 volts, and is therefore two orders of magnitude less than for a portable thermionic receiver. Transistors are also being applied to cathode-ray oscillographs.[30]



(*a*)



(*b*)

Fig. 2.—Typical three-stage amplifier circuits.
(*a*) Hearing-aid amplifier.
(*b*) Servo-amplifier.

Transistor amplifiers for analogue computers have been described by Ettinger.[31] One type is a bi-directional voltage discriminator which operates a centre-stable relay in response to an input signal change of 10 mV. The input resistance is 50 000 ohms and the amplifier power gain is about 32 dB. Drift either of voltage or current may be a problem in simple circuit arrangements of this nature. More elaborate circuits described by Blecher[11] include a summing amplifier with a d.c. accuracy of better than one part in 5 000 over the range 0–50° C and an integrator that generates a voltage ramp linear to within one part in 8 000.

A 2-watt servo-amplifier circuit based on silicon transistors is shown[35] in Fig. 2(*b*). The amplifier drives the control phase of a 3 400-ohm 2-phase a.c. servo-motor from a 400 c/s input signal. The voltage gain is stabilized at 400 by negative feedback through the resistors R_f and R_e. The nominal ambient operating temperature is 71° C, and 100° C is quoted as the maximum permissible ambient temperature.

In a recent paper on transistor amplifiers for radar video circuits, it is stated that, although silicon transistors will operate without failure up to 150° C, the parameters can change by as much at 50% as the upper temperature limit is approached.[36] At present it appears that operation under ambient temperatures of 100° C is about the best that can be achieved with silicon transistors.

When the output power required in a servo mechanism is in excess of that obtainable from readily available power transistors,[33, 34] the transistor amplifier may be followed by a magnetic-amplifier output stage. Such amplifiers are described by Weyrick[37] and by Zeller.[38] The servo input signal may be 400 or 60 c/s, and it is usually found convenient to demodulate at the output of the transistor amplifier before feeding into the magnetic amplifier. Certain magnetic-amplifier circuits respond to a.c. input signals, but conversion to direct current at this stage provides a convenient point for the insertion of d.c. shaping networks to give suitable gain and phase response in systems.

Transistors are finding numerous uses in instrumentation and a few references to this class of application are given.[41–44]

In many of the References, and in a few of the circuit diagrams in the present paper, transistors are designated by type number,

and in this connection the information on units given by Schwartz is useful.[45] The transistor symbols used in the paper were originally devised for point-contact units, but are now also used for junction transistors.

(3) TRANSISTOR REGULATOR CIRCUITS

(3.1) Voltage or Current Regulation

Transistors may be used for voltage or current regulation of d.c. power-supply systems.[46–52] Shunt or series regulating systems may be used, and the most convenient reference device is an avalanche diode. This is a silicon diode with a reverse breakdown voltage that is almost constant with change of reverse current over a range, such as 10–30 mA. Good reference diodes at present available have working voltages of about 5 volts, reverse slope resistances of 10–20 ohms and temperature coefficients of less than 0·01% per deg C. Higher-voltage diodes tend to be of greater slope resistance and to have inferior temperature stability.

A typical series voltage-regulating circuit is shown in Fig. 3(*a*).

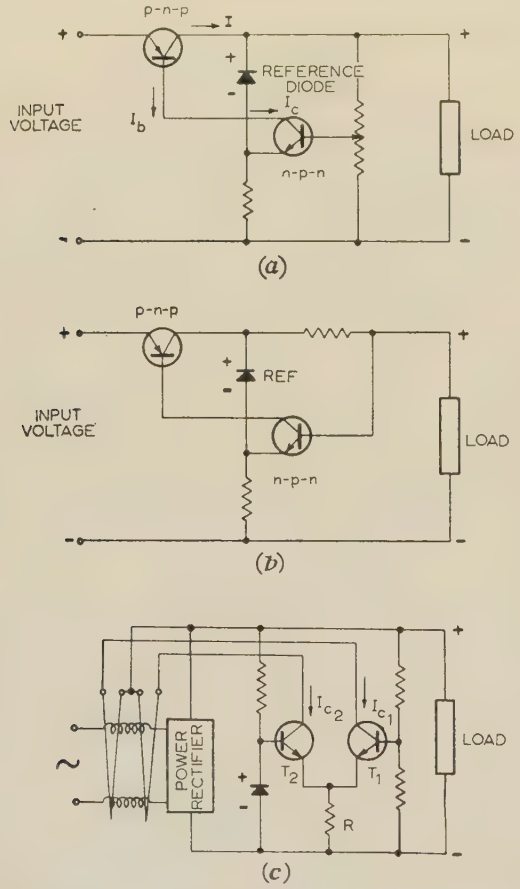


Fig. 3.—Voltage or current regulation of d.c. load supply sources.
(*a*) Basic circuit for voltage regulation.
(*b*) Basic circuit for current regulation.
(*c*) Voltage-regulated rectifier with magnetic amplifier and push-pull transistor control.

The voltage-sensing bridge formed by the reference diode and the resistors is so arranged that, if the load voltage increases, the base of the *n-p-n* unit becomes less positive with respect to the emitter, and so the base and collector currents decrease. The collector current of the driver transistor is the base current of the *p-n-p* power unit, and so the overall effect is a decrease

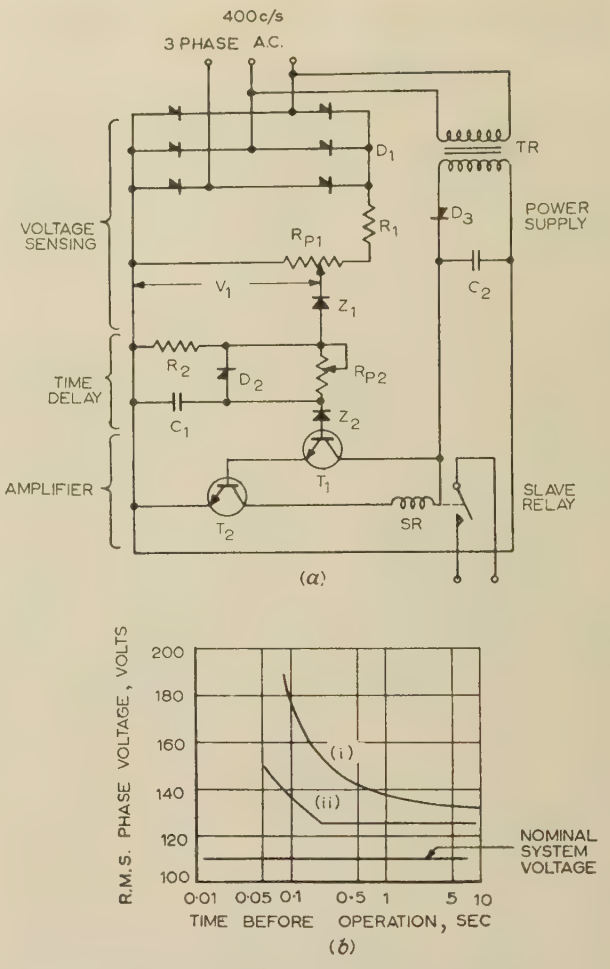


Fig. 4.—Over-voltage relay using transistors.
(*a*) Circuit diagram.
(*b*) Voltage/time characteristics.
(i) Upper limit of adjustment.
(ii) Lower limit of adjustment.

of load current and therefore regulation of the output voltage. Where load-current regulation is required, the circuit may be rearranged as shown in Fig. 3(*b*).

The accuracy of regulation depends on the performance of the reference diode and on the system loop gain. Typically, if ±1% voltage regulation is required over a temperature range such as −55° C to +55° C for an input voltage variation of, say, 25–29 volts, a three- or four-stage transistor circuit is required for adequate gain. The power rating and dissipation required of the output transistor are high, but when suitable units are available, transistor regulators may be expected to replace carbon-pile systems, with improvements in accuracy and reliability. In some applications carbon piles may be used as transistor-controlled output stages. Thyratrons also have been used as output regulating elements in a.c. to d.c. regulators.

A transistor-controlled magnetic-amplifier-regulated rectifier system proposed by Chase is shown in Fig. 3(*c*). The push-pull transistor arrangement gives improved temperature stability, since variations of I_{c1} and I_{c2} with temperature cancel in the windings of the magnetic amplifier and produce no control effect. Increase in the load voltage increases the output of the transistor T_1, and through the coupling resistor R this results in a decrease in I_{c2}; the difference current I_{c1}–I_{c2} controls the magnetic amplifier and decreases the load current to achieve regulation.

The design of transistor regulator circuits is simplified if both

p-n-p and *n-p-n* units are available, since coupling between them is straightforward.[4]

(3.2) Transistorized Relay Circuits

Transistors may be applied to various types of protective and warning relay systems that may be considered as forms of regulating circuits.

An over-voltage relay circuit, devised by Schuh,[53] is shown in Fig. 4. At normal system voltage, V_1 is below the breakdown voltage of the reference diode Z_1 and the circuit is inactive with C_1 virtually uncharged. At a preset level of system over-voltage, the diode Z_1 conducts and charges C_1 through the potentiometer R_{P2}. As C_1 charges, a second reference diode Z_2 begins to conduct and base current flows into the transistor T_1, which controls T_2 and begins to energize the slave relay coil. When the charging of C_1 has continued for a certain period of time the current in the relay coil is sufficient to cause operation. But if, at any time before the relay operates, the system voltage returns to normal, C_1 discharges through the circuit D_2, R_2, thus, in effect, resetting the whole circuit. The system voltage/time characteristics are adjustable between the limits (i) and (ii) in Fig. 4(*b*) by decrease of R_{P2}. The complete relay when packaged for aircraft use occupies 2 in × 3 in × 4 in and weighs less than 1 lb.

A negative-phase-sequence alarm relay in the United States involves transistor control and can be set to pick up at a value of negative-phase-sequence current of only 7% of the rated current. This represents 3½ times the sensitivity of previous relay alarm systems and detects unbalance load conditions before adverse effects on the system generators can occur.[54]

The application of transistors to power-system-protection relays has also been considered by Adamson and Wedepohl,[55, 56] and others.[57] Certain circuit design problems of susceptibility to d.c. transient conditions are involved, but transistors are of practical value in the design and development of new protective-gear systems and relays.

(4) SWITCHING-TRANSISTOR POWER-SUPPLY UNITS

The concept of transistor operation as a switch element, as examined by Bright,[58] and others,[59, 60] has proved of considerable importance. With suitable base drive conditions the emitter-collector impedance of a transistor can be varied from, say, 100 kilohms (the open-switch condition) to less than 1 ohm (the closed-switch condition) in times of the order of $1/f_\alpha$, where f_α is the common-base α cut-off frequency. The extreme conditions do not involve large power loss, since either the current or voltage of the transistor is small. Thus switched modes of transistor operation are of large power-handling capacity and of high efficiency.

(4.1) Saturable-Transformer Square-Wave Oscillators

A two-transistor square-wave oscillator, controlled in frequency by the saturation flux limits of the output transformer core,[61, 62] is shown in Fig. 5(*a*). The half-cycle is considered when transistor A is conducting and the battery voltage V is applied to winding 2. This voltage causes flux increase in the transformer core and induces a voltage across winding 1 which provides base current for the transistor. During the period of flux increase, the induced voltage in winding 4 is in the sense that maintains transistor B non-conducting, since the base is positive with respect to the emitter. Because of the induced voltage in winding 3, the total emitter-collector voltage appearing across transistor B in the cut-off condition is equal to twice the battery voltage.

At the end of the half-cycle the flux in the transformer core

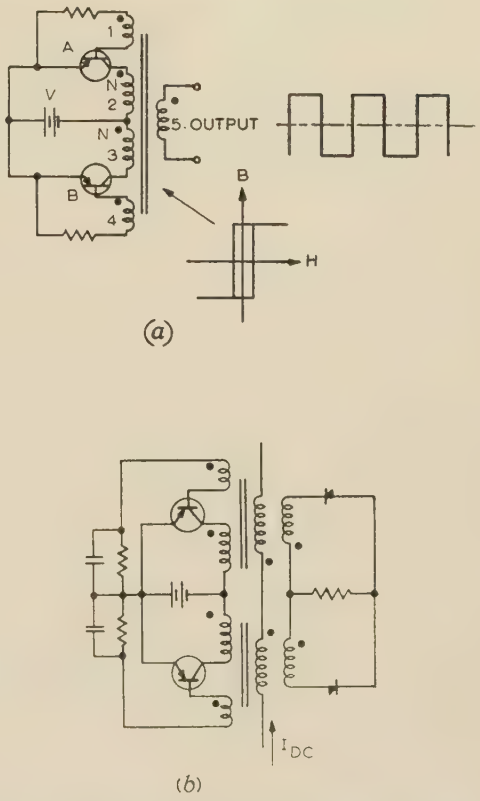


Fig. 5.—Saturable-transformer square-wave oscillators.
(*a*) Basic oscillator.
(*b*) Multivibrator of variable frequency derived from the basic circuit.

reaches the positive saturation level, $+\Phi_s$, and the drive and hold-off voltages in the base circuits of transistors A and B then collapse. This permits the two transistors to interchange action so that the input voltage is applied to winding 3 and the core flux then proceeds towards negative saturation. If windings 2 and 3 have N turns each, the voltage-flux relationship is

$$\int_0^{\frac{1}{2}f} V dt = \int_{-\Phi_s}^{+\Phi_s} \frac{N}{10^8} d\phi \quad . \quad . \quad . \quad . \quad . \quad (6)$$

whence the frequency of oscillation is given by

$$f = \frac{V 10^8}{4N\Phi_s} \quad . \quad . \quad . \quad . \quad . \quad . \quad (7)$$

Thus, for a given saturable transformer the frequency of the alternating output is proportional to the input voltage. If the input voltage is increased, the output voltage increases proportionately, but the period of oscillation decreases (i.e. the frequency increases) and the voltage-time integral of the output is unchanged.

With currently-available 3 amp transistors, output powers in excess of 50 watts have been obtained at efficiencies of over 80% and in ambient temperatures of 70° C. Switching frequencies from 100 c/s to 10 kc/s are commonly used.

Such transistor convertors are attractive as supply sources for magnetic amplifiers, a.c. servo motors and other components used in low-power servo-control systems. For example, in airborne military applications, transistor power supplies may replace small rotating machine inverters (which, at present, produce 400 c/s supplies from the aircraft battery system), with advantages in reliability and weight. For magnetic

amplifiers, a supply source of constant voltage-time integral is advantageous, since it corresponds to constant excitation of the cores in terms of flux swing, and the rectangular waveform of the convertor output is acceptable.[63] Other applications of the circuit as transducers in telemetry systems and as integrators in analogue-to-digital conversion are described by Royer.[61]

In Fig. 5(*b*) two cores with d.c. bias have been substituted for the single core of the basic circuit, so that the frequency of switching no longer depends on knee-to-knee saturation flux swings but is varied by change of bias. Frequency control over a range of more than 10 : 1 is possible.[64] Oscillator action has been investigated by Chen[65] and Mogen,[66] but in some aspects further study is perhaps required.

(4.2) Phase Locking of Square-Wave Oscillators

Circuits for the phase-locking of square-wave transistor oscillators appear desirable in applications such as the control of 2-phase a.c. servo motors, since normal methods of obtaining a phase shift in a sinusoidal supply system, e.g. by a capacitor in series with one phase winding of the motor, are not applicable.

A technique for accomplishing 90° phase-locking of two square-wave oscillators has been described in detail elsewhere.[67] The phase difference of the two waveforms is detected by considering the sum and difference of the output voltages in two auxiliary circuits containing saturable-core reactors. If the phase relationship is in error, voltage becomes available in one or other of the auxiliary circuits for correction of the relative switching moments of the two oscillators.

By suitable design of the saturation flux swing of the auxiliary saturable-core reactors, the circuit may be arranged to provide any desired angle of phase lock. Hence three convertors may be set in 120° phase sequence to give a 3-phase system if required. These techniques for linking square-wave oscillators to form polyphase systems are likely to enhance their attractiveness for power-supply purposes.

(4.3) Voltage Step-up D.C. Transistor Power Units

Thermionic circuits needing direct-voltage h.t. supplies derived from a low-voltage supply battery are, at present, dependent on the use of rotary convertors or vibrating-reed 'chopper' power units. Switching transistor convertors are tending to supersede these methods because of high efficiency and the superior reliability inherent in the absence of moving parts and contacts.

The transistor-convertor circuit shown in Fig. 6(*a*) is a simple transformer-coupled relaxation oscillator in which energy is stored in the inductance of the transformer during the 'on' period of the transistor and is delivered to the output circuit at a higher voltage during the 'off' period. The circuit action has been discussed by Light and Hooker,[68] who give details of a number of convertors having efficiencies of 70–75% and outputs of up to 2 watts. Applications to radiation monitors have been described by Franklin and James.[69]

For higher-power units the circuit of Fig. 6(*b*) is more suitable and gives efficiencies of over 80%. This arrangement is simply the square-wave-oscillator circuit of Fig. 5(*a*) followed by full-wave bridge rectification of the output voltage, which is stepped-up by the transformer. The resistance *R* ensures that the potential of P is suitably negative during the initial switch-on condition, and the diode D carries the base currents once normal running is established. Without some arrangement of this nature the convertor may not start on full load. Some added inductance at the point X in Fig. 6(*b*) may also be helpful in permitting starting with a heavily smoothed load.

Typically a 50-watt unit, switching at about 700 c/s, measures 4 in × 3 in × 2½ in and weighs approximately 1 lb. Recent higher-power designs include a 175 watt 70° C (ambient) unit using selected transistors.[70]

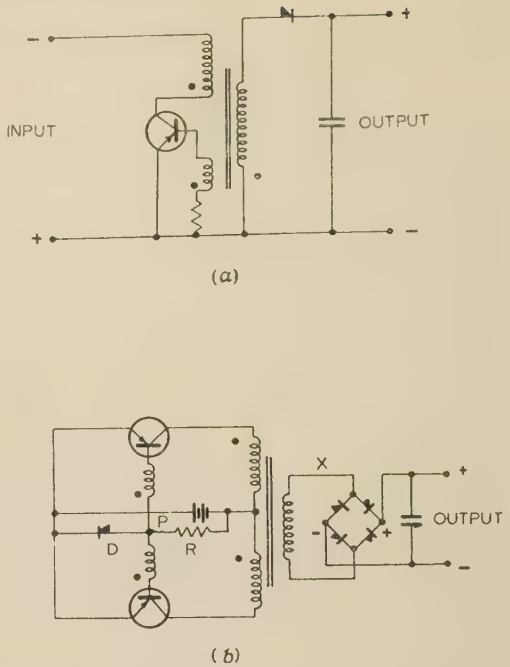


Fig. 6.—D.C.-convertor circuits for voltage step-up.
(*a*) Relaxation-oscillator conversion circuit.
(*b*) Typical convertor circuit based on the saturating-transformer oscillator.

(5) SINUSOIDAL JUNCTION-TRANSISTOR OSCILLATORS

(5.1) Basic Feedback-Oscillator Circuits

Junction-transistor amplifying connections may be made to oscillate by external feedback from the output to the input. The signal at the input terminals must be of suitable phase and amplitude to give an output which, returned through the feedback circuit, sustains the input. In other words, at the oscillation frequency the phase shift round the complete loop must be zero and the loop gain must not be less than unity. The maximum frequency of oscillation of a transistor is normally several times the α cut-off frequency, but this is an extreme condition for self-oscillation, and the upper usable limit is less by perhaps an order of magnitude. The bilateral nature of a transistor, owing to internal feedback between output and input, complicates the analysis of transistor oscillators, particularly at high oscillation frequencies, and the equivalent circuits used must be suitably chosen. In general, the frequency stabilities observed with change of supply voltage, load and temperature are of the order of one part in 10^3.

Four typical feedback oscillator circuits are given in Fig. 7 to illustrate the arrangements used. In Fig. 7(*a*) a tuned collector load is used, and there is transformer feedback into the base circuit. Hartley and Colpitts type arrangements are shown in Figs. 7(*b*) and 7(*c*), and a quartz-crystal-controlled circuit is given in Fig. 7(*d*). This latter circuit has a Colpitts-type configuration, together with a tuned collector load, and is stated to have excellent characteristics.[3]

For low-frequency oscillator circuits resistance-capacitance feedback networks may be used, as described by Hooper and Jackets.[71] The frequency range covered is from a fraction of a cycle per second up to some tens of kilocycles per second.

Design considerations for junction-transistor oscillators have been treated by Oakes,[72] and illustrated by a detailed design of a class-B circuit.

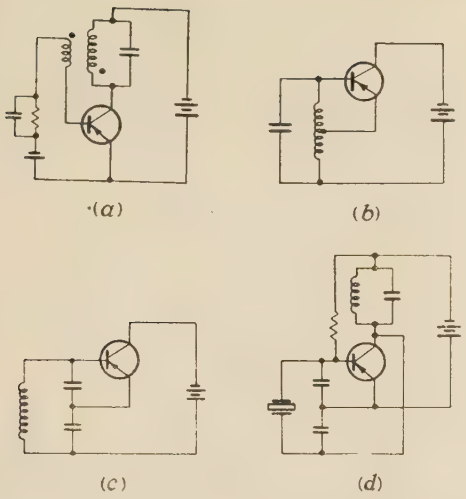


Fig. 7.—Some junction-transistor oscillator circuits.

(*a*) Oscillator with transformer feedback from the collector tuned circuit to the base.
(*b*) Hartley oscillator arrangement.
(*c*) Colpitts oscillator arrangement.
(*d*) Quartz-crystal-controlled oscillator.

(5.2) Two High-Grade Oscillator Arrangements

A frequency-stable circuit devised by Keonjian[73] is given in Fig. 8(*a*). The basic circuit, centred around transistor 1, is a Colpitts arrangement in which the tuning inductance is replaced by a series inductance-capacitance circuit. This is found to give good stability and to be non-critical in adjustment. The effect of supply-voltage variation on the operating-point bias, and hence on the frequency of oscillation, is reduced by the auxiliary transistor 2 and the voltage-divider circuit R_{11}–R_{12}. To compensate for ambient-temperature effects it is necessary to decrease the emitter current of transistor 1 as the temperature rises, and this is done by making R_3 a temperature-sensitive resistance with a negative temperature coefficient, namely a thermistor. The performance achieved in a 1–2 Mc/s circuit is a frequency stability of 10–20 parts in 10^6 for changes in supply voltage of 10% and for an ambient temperature range of 10 or 20° C.

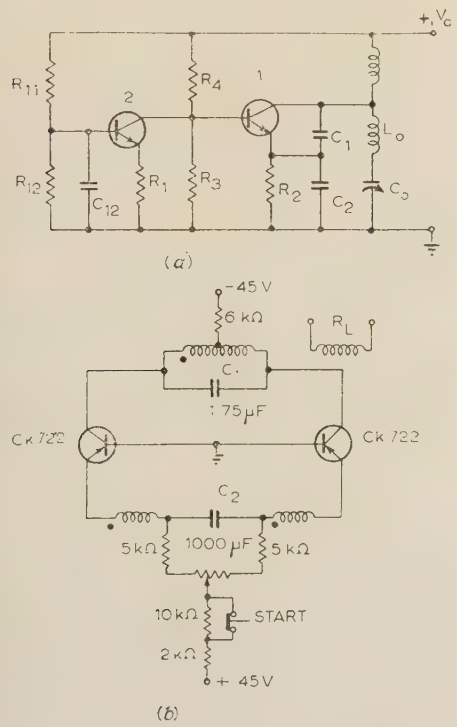


Fig. 8.—Special-performance sinusoidal oscillators.

(*a*) Oscillator with frequency stability of one part in 10^5.
(*b*) Balanced oscillator with low second-harmonic distortion (0·006%).

The oscillator circuit shown in Fig. 8(*b*) is of low second-harmonic distortion and is an adaptation by Grim[74] of a thermionic-valve circuit designed by Noble and Baxandall[75] to supply a second-harmonic type of magnetic modulator. The transistors are operated as switches to give stability independent of temperature-sensitive parameters, and the conduction periods are made equal by circuit symmetry controlled by the large capacitance C_2 and the associated resistances. A 400 c/s model has a power output of 20 mW at an efficiency of 3%, but the second-harmonic distortion remains less than 0·006% over a wide range of test operating conditions, including a 20° C change of temperature.

(5.3) Three-Phase Power Oscillator Circuits

Sinusoidal 3-phase transistor oscillator systems have been investigated recently, and a number of satisfactory circuit arrangements have been devised.

In a sinusoidal system, phase locking may be based on the principle that the sum of two phase voltages is instantaneously

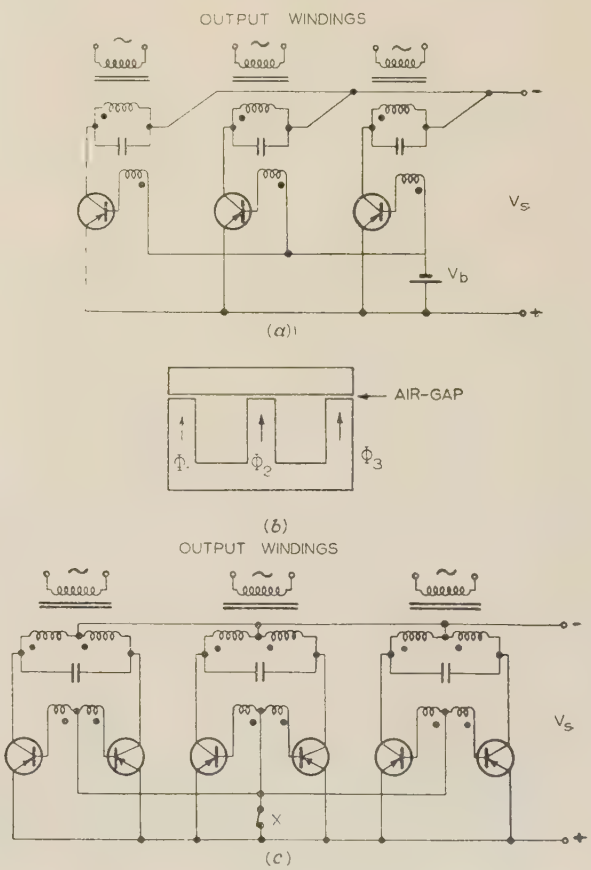


Fig. 9.—Three-phase sinusoidal oscillator circuits.

(*a*) Class-A circuit.
(*b*) Three-legged core used for tuned transformer.
(*c*) Class-B push-pull circuit.

equal, but opposite, to the remaining phase voltage. It may be arranged, therefore, for each of three transistor oscillators to be driven by the sum of the voltage outputs of the two other oscillators—this sum being derived either by resistance networks or from extra windings on the tuned-circuit chokes.

A 3-phase class-A oscillator arrangement is shown in Fig. 9(*a*), in which the coils of the tuned circuits are wound on separate limbs of a symmetrical 3-phase core such as that shown in Fig. 9(*b*).[76] The magnetic fluxes in the three core limbs necessarily sum to zero, and therefore the tuned-circuit voltages must also sum to zero, which is the condition for a balanced 3-phase system. The theoretical r.m.s. voltage across the tuned circuits for full output swing is $V_s/\sqrt{2}$, and the required collector-emitter rating of the transistor is $2V_s$. The battery voltage V_b provides d.c. bias in the base circuits to set the class-A operating point. The tuned circuits should be designed to have circulating currents of several times the load currents. Power outputs of a few watts per phase at efficiencies of 40% have been obtained with this circuit arrangement.

For class-B push-pull operation the circuit shown in Fig. 9(*c*) has been devised, and is a more practical arrangement than the class-A circuit because of greater power-handling capacity and efficiency. A 115-volt 400 c/s development model for 20-volt battery input gave 20 watts of power per phase with an overall circuit efficiency in the range 50–60%. The frequency variation from zero to full load was ±5%, and the output voltage decreased from 150 to 115 volts with load variation (corresponding to an output impedance of about 100 ohms). Starting is accomplished by a simple pushbutton arrangement, which opens and closes the switch X in the common base lead before the application of load. The circuit will also start with the diode-biasing method shown in Fig. 6(*b*). The phase sequence of the oscillation can be controlled by a simple detector circuit. This may take the form of a capacitive circuit for advancing one phase voltage by 60°, which when added to a second phase voltage gives either zero or a substantial output voltage, depending on the phase rotation.

(6) LOW-LEVEL SWITCHING CIRCUITS

(6.1) Transistor 'Choppers' or Modulators

Transistors can perform the functions of switches down to signal levels of a fraction of a millivolt, and germanium units may be used over the temperature range −50° C to +90° C in this class of operation. The absence of moving parts gives an advantage over mechanical 'choppers', and transistor 'choppers' have practically zero phase shift up to frequencies of 10 kc/s or higher.

The characteristics of alloy-type transistors have been studied by Bright[77] and Kruper[78] with respect to low-signal switching performance, and for non-symmetrical units the inverted condition of operation has been shown to be decidedly superior to the normal condition [see Figs. 10(*a*) and 10(*b*)]. It is found that the transistor base-switching drive produces a certain minimum collector-emitter voltage that feeds through in the 'closed' switch condition and a certain minimum leakage current that exists in the 'open' switch condition. The feed-through voltage may be 10–40 mV in the normal connection and less than 1 mV in the inverted connection, and the leakage current may decrease from 40 to 1 μA.

The feed-through voltage may be further reduced by the use of a balanced-circuit arrangement, as shown in Fig. 10(*c*). The base drive is a square wave of about 1-volt amplitude. On one half-cycle transistor 1 acts as a closed switch and transistor 2 is open—hence the signal is applied to the output transformer—and on the second half-cycle the transistor conditions are reversed.

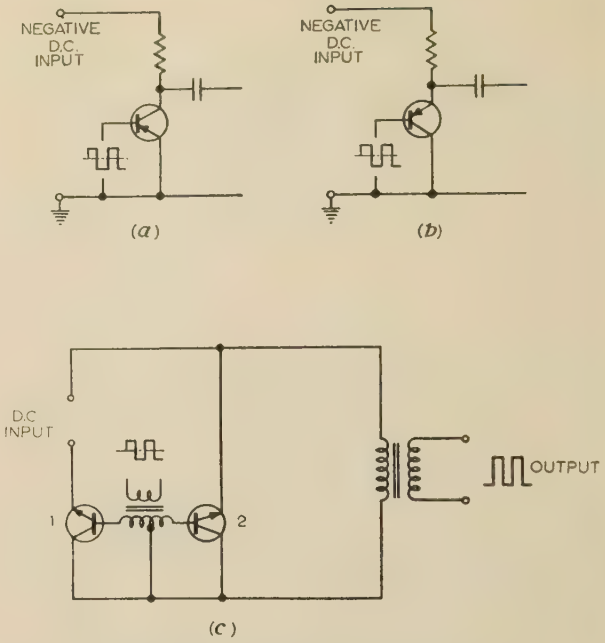


Fig. 10.—Transistor chopping circuits.
(*a*) Basic circuit with *p-n-p* transistor.
(*b*) Basic circuit with inverted operation of the *p-n-p* transistor.
(*c*) Balanced low-level chopper circuit for d.c. to a.c. signal conversion.

The transistor feed-through voltages in the absence of signal are applied to the transformer primary winding in a unidirectional sense and so give no a.c. output if exactly matched. In practice, matching appears to be possible to within 0·1 mV for a wide temperature range, and so a signal level down to a fraction of a millivolt can be handled. The upper signal level should not exceed the amplitude of the base switching voltage, if open-switch conditions are to be obtained. The input impedance is determined basically by the load resistance and is normally made low for thermocouple-type signals.

For impedance levels of a few thousand ohms to hundreds of thousands of ohms, the silicon-diode modulator circuit described by Moody for d.c. to a.c. conversion is of interest.[79] At an input-impedance level of 300 kilohms the circuit is sensitive to currents of 10^{-8} amp and the zero stability is of the order of 10^{-10} watt. The frequency range extends from audio frequency to at least 100 kc/s, and operation is considered acceptable up to 80° C.

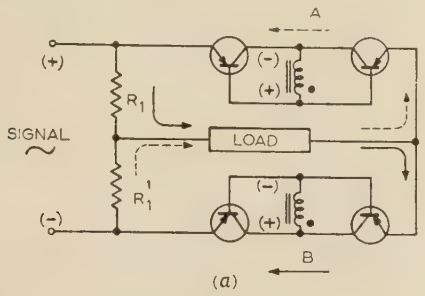
(6.2) Demodulators or Phase-Sensitive Rectifiers

A *p-n-p* (or *n-p-n*) transistor is of high impedance, between collector and emitter, only if the base is positive in potential (or negative for *n-p-n* units) with respect to both of the other terminals. It follows that, for switch action in an a.c. circuit, where the voltage to be switched is greater than the base drive voltage, two transistors must be used in a back-to-back configuration, as described by Bright.[58]

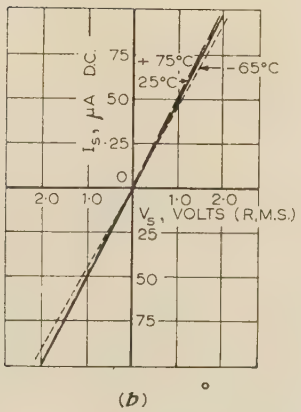
A four-transistor demodulator comprising two such switch arms is shown in Fig. 11(*a*). The base drive voltages, which may be either square waves or sine waves, determine the phase reference, and the signal should be of the same frequency and phase. For the polarities indicated on the diagram the switch arm A is cut off and B is on, and so the current flow is as shown by the full-line arrows. In the next half-cycle the polarities of signal and base voltages are reversed and the transistor arms interchange functions, but the current flow is unidirectional

through the load as indicated by the dotted-line arrows. With reversal of signal phase relative to the base drive voltages the output current direction reverses, thus giving the circuit phase-sensing properties.

Characteristics obtained from such a demodulator circuit by Decker[80] are given in Fig. 11(*b*). The 400 c/s signal-voltage



(*a*)



(*b*)

Fig. 11.—Phase-sensitive rectifier using four transistors.
(*a*) Circuit arrangement.
(*b*) Typical performance characteristics.

range claimed is from 20 mV to 30 volts (r.m.s.) and the drift is less than 10^{-8} watt for the temperature range $-65°$ C to $+75°$ C with germanium transistors. The input impedance level is determined mainly by the resistors R_1, R_1' and is 10 kilohms. Typical applications have been in the servo-mechanism field with frequencies of switching of 50–2400 c/s, but the upper limit could be tens of kilocycles if desired. The lower limit of signal sensitivity, based on the performance of transistor choppers, should be below 1 mV.

In certain applications where the temperature range is not wide and the signal power levels involved are not particularly low, demodulator circuits may require less than four transistors. Many circuit configurations are possible—Sutcliffe, for example, has recently described a two-transistor half-wave demodulator in which the transistors function both as linear signal amplifiers and as non-linear rectifying elements.[81]

(6.3) Transistor and Magnetic-Core Analogue Computation Circuits

Rectangular-loop magnetic cores and transistors used as switches may be combined to form circuits capable of analogue computation.[82–85] The principles of operation involved are most conveniently discussed with reference to a circuit having a specific function.

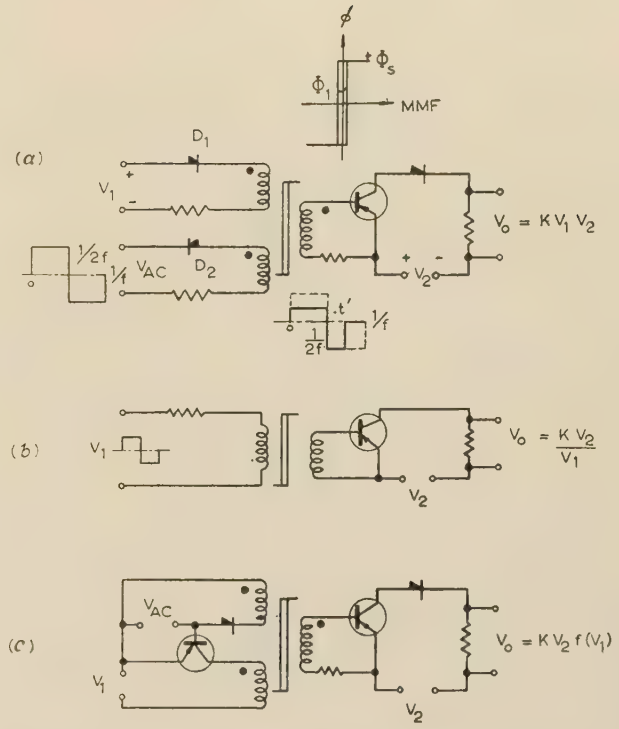


Fig. 12.—Transistor and magnetic-core analogue computation circuits.
(*a*) Single-quadrant multiplying circuit.
(*b*) Single-quadrant dividing circuit.
(*c*) Basic transcendental function computer circuit.

The action of the multiplying circuit given in Fig. 12(*a*) is considered. This has a rectangular drive voltage V_{AC} and gives a mean output voltage V_0 proportional to the product of two direct signal voltages V_1 and V_2. In the half-cycle when the diode D_2 blocks the drive voltage from the circuit, the signal voltage V_1 resets the core flux from $+\Phi_s$ to Φ_1 according to the relation

$$\left.\begin{aligned}\Phi_s - \Phi_1 &= \int_0^{\frac{1}{2f}} V_1 dt \\ &= \frac{V_1}{2f}\end{aligned}\right\} \quad \ldots\ldots \quad (8)$$

In the next half-cycle the drive voltage returns the flux to Φ_s, but because $V_{AC} > V_1$, the operation is completed in a period $1/(2f)$ to t', where

$$V_{AC}\left(t' - \frac{1}{2f}\right) = \Phi_s - \Phi_1 \quad \ldots\ldots \quad (9)$$

and so, from eqns. (8) and (9),

$$t' - \frac{1}{2f} = \frac{V_1}{V_{AC}} \frac{1}{2f} \quad \ldots\ldots \quad (10)$$

While the flux is changing from Φ_1 to Φ_s the base-circuit voltage bottoms the transistor and the second signal V_2 appears as output voltage. The mean output voltage over a complete cycle is therefore

$$V_0 = \frac{V_2\left(t' - \frac{1}{2f}\right)}{1/f} \quad \ldots\ldots \quad (11)$$

whence, from eqn. (10),

$$V_0 = \frac{V_1 V_2}{2V_{AC}} \quad . \quad . \quad . \quad . \quad . \quad . \quad (12)$$

Thus the mean output voltage depends on the product of the two signal voltages. In the mode of action described, V_1 and V_2 are assumed to be direct voltages, but this is not essential provided that they are suitably related to the drive voltage. With a four-quadrant version of this circuit handling voltages of 5–50 volts, Van Allen[82] reports an overall accuracy of about $\pm 2\%$. Switching frequencies of up to 5 kc/s have been successfully used.

A single-quadrant dividing circuit based on similar principles is shown in Fig. 12(*b*). The signal voltage V_1 must be sufficient to swing the core flux from one saturation limit to the other. The time taken to accomplish this, which is also the conduction period of the transistor, decreases as the voltage V_1 is increased, and hence a mean output voltage is obtained proportional to V_2/V_1. With a four-quadrant circuit arrangement using signal voltages in the range 2–40 volts, an accuracy of $\pm 2\frac{1}{2}\%$ has been achieved.[83]

With the basic circuit shown in Fig. 12(*c*), a considerable number of analogue operations may be performed depending on the waveforms of the applied voltages V_2 and V_{AC}. Schaefer and Van Allen present results for operations such as obtaining an output voltage proportional to the square root and other fractional powers of an input voltage, the arc sine of an input voltage and the sine.[84] The underlying principle throughout is that of width and amplitude modulation of voltages of specific waveshapes, using the flux changes in a magnetic core to store voltage-time relations for a half-cycle of the operating frequency. The cores need to be of high remanence and high permeability, and therefore square-loop nickel-iron tape cores are normally used.

The possibilities of devising circuits for analogue-function operations without magnetic cores have been briefly mentioned by Schiewe and Chen.[84]

(6.4) Transistor Digital-Computer Circuits

In the digital-computer field attention is being given to the use of transistors in order to reduce bulk and power consumption and to increase reliability. Two-state circuits using point-contact transistors have been used, but recent trends have been towards junction-transistor arrangements.[86] A complete survey of these circuits is not possible here, but two recent transistor-ferrite-core arrangements are shown in Fig. 13.

The basic circuit in Fig. 13(*a*) shows transistor coupling between two magnetic cores of square-loop ferrite material (typically only 2 mm in diameter) carrying binary notation information depending on the direction of remanent magnetization.[87] The transistor coupling makes it possible for the magnetic state of core 1 to be transmitted to core 2 on the application of a drive pulse to the windings Y. With a stabilizing bias this circuit is capable of operating at temperatures of up to 100° C with germanium transistors.

A single-core counter arrangement is shown[88] in Fig. 13(*b*). By means of the network C, R_1 and R_2, the input current, which is a unidirectional pulse, is converted into two successive pulses of opposite polarity that pass through winding *a*. The first of these pulses changes flux in the core in a direction which turns the transistor on, and the collector current passing through the feedback winding *b* maintains the change of flux and masks the second pulse in winding *a*. When the next input pulse appears the first pulse in winding *a* produces no change of flux and the second switches the core back to its initial state.

The circuits of Fig. 13 have been operated at input pulse rates of up to 2×10^5 pulses/sec with transistors having α cut-off

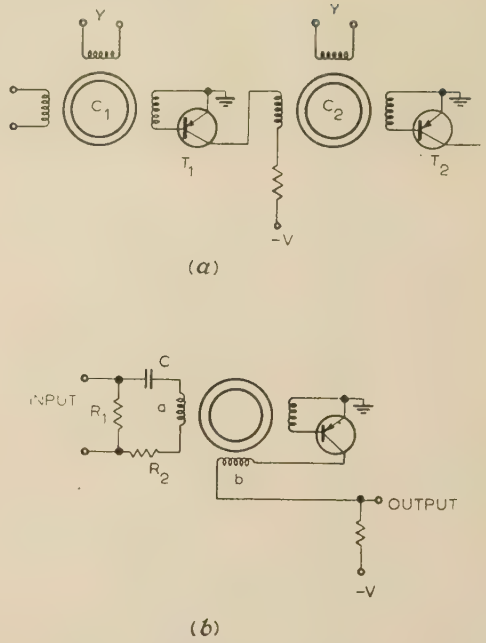


Fig. 13.—Transistor and ferrite-core digital-computer circuits.
(*a*) Basic circuit for shift register.
(*b*) Single-core counter arrangement.

frequencies of less than 1 Mc/s. In a data transmission application the use of this type of circuit in place of thermionic-valve circuits effected a 3000 : 1 reduction in power consumption (150 mW against 500 watts) and at least a 100 : 1 reduction in volume.

A rather different arrangement for counting using a series of flux-storage levels in a magnetic core is the basis of a transistorized time-delay device described by Pittman.[89]

(7) WAVEFORM SHAPING AND GENERATION

Many techniques of waveform shaping and generation are applied in thermionic-valve circuit design, and similar techniques are possible in transistor circuit work. The circuit principles employed are often much the same for thermionic-valve and transistor arrangements. Among the techniques to be mentioned are integration and differentiation, multivibrators and bistable circuits, blocking oscillators, sawtooth waveform generation and switching circuits for scaling.

(7.1) Integration and Differentiation

An amplifier configuration with output voltage v_0 and capacitive feedback between output and input is considered. The feedback current is given by Cdv_0/dt, and if the amplifier is of high gain the overall effect will be to produce an output that is the integral of the input voltage. Fig. 14(*a*) shows the technique applied to a simple transistor arrangement for the integration of a 1000 c/s square-wave input.[62]

For purposes of differentiation the amplifier feedback loop must provide integration, and this can be obtained by applying to a capacitor a feedback current derived from the output voltage through a large feedback resistance. The voltage across the capacitor then contains a term proportional to the integral of the output voltage. Fig. 14(*b*) shows a specific circuit arrangement given by Oakes for the differentiation of a 1000 c/s triangular waveform.[90] With an input of 40 volts (peak-to-peak), the square-wave output obtained was just over 6 volts (peak to-peak). The circuits [Figs. 14(*a*) and 14(*b*)] require stabilization

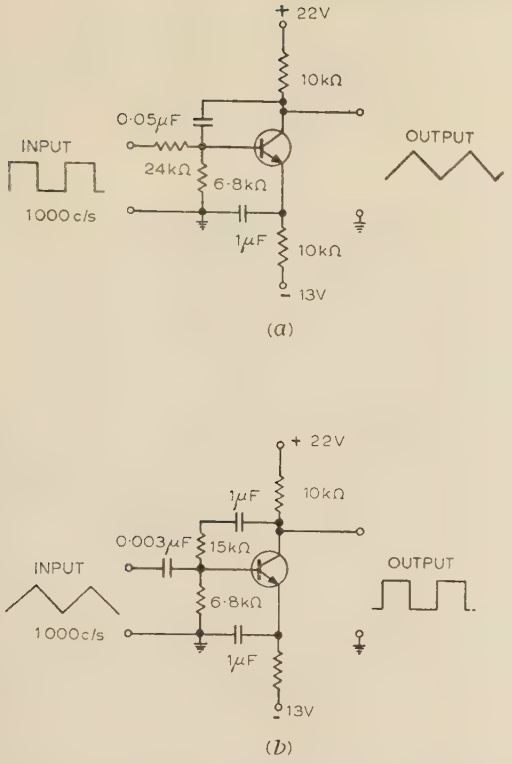


Fig. 14.—Integration and differentiation of waveforms.
(*a*) Integrating circuit.
(*b*) Differentiating circuit.

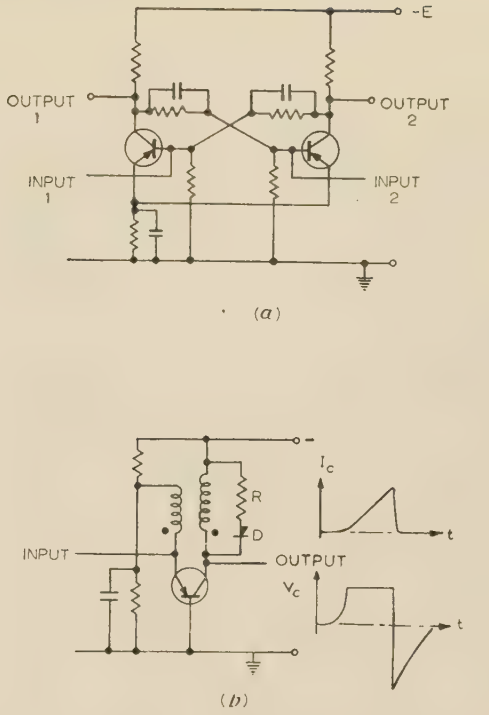


Fig. 15.—Flip-flop and blocking-oscillator circuits.
(*a*) Flip-flop multivibrator circuit.
(*b*) Relaxation blocking oscillator.

of the operating point for temperature and supply-voltage variations if low-frequency precision operation is required.

(7.2) Multivibrators and Blocking Oscillators

Junction-transistor multivibrator circuits—monostable, bistable or free-running—have been given by Prugh[91] and others.[93, 94, 95] Fig. 15(*a*) shows a typical flip-flop arrangement which depends for its action on cross-feedback between collector and base and so is similar to Eccles–Jordan arrangements in thermionic-valve circuits.

Minority carrier storage may represent a limitation in switching speed of transistor flip-flop and related circuits. The phenomenon occurs if a transistor is over-driven (bottomed) by a base current in excess of the collector-current requirement, since the base region becomes flooded with minority charge-carriers which have to be dispersed for the transistor to resume normal-region operation. A study of this effect has been made by Moll[92] in respect of large-signal switching behaviour of junction transistors. If ω_0 is the α cut-off angular frequency of a transistor, Moll shows that, for the common-emitter connection, base currents of about $\frac{1}{3}I_c$ result in current turn-on times of approximately $3/\omega_0$, and a similar reverse turn-off current results in storage times of approximately $5/\omega_0$ to $10/\omega_0$ followed by decay times of about $3/\omega_0$. Reduction of turn-on or turn-off time to $1/\omega_0$ requires drive currents about equal to the load current. These typical figures imply that for fast circuit action the transistors should not be allowed to bottom and that the base drive should be large in the normal region of operation. From Moll's analysis it may be noted that carrier storage in a non-symmetrical transistor is mainly related to the inverse characteristics ω_I and α_I.

Linvill[93] has described a flip-flop circuit in which breakdown diodes are used to prevent the transistor collector voltages from becoming zero by limiting the range of voltage swing of the transient. Since bottoming does not occur, the switching times achieved are as low as several times $1/\omega_0$, and a binary counter operating at 1·25 Mc/s is described. The pulse requirements for switching the counter are analysed and it is shown that the switching charge, which is the significant quantity, should be about $1/\omega_0$ times the difference in conduction currents of the two transistors in the stable state.

If two-transistor monostable circuits depend on capacitance timing, there is difficulty in obtaining suitable switching speed to give short pulses.[102] Blocking-oscillator arrangements, which depend on inductance for pulse-length timing, tend to be somewhat better in this respect. There are many possible circuit configurations, of which a typical arrangement is shown in Fig. 15(*b*). The collector current is initially low, but when the trigger input pulse is applied the feedback loop causes the transistor to bottom and the battery voltage appears across the magnetizing inductance of the transformer winding, which causes a linear rise of current until I_c equals the value permitted by the emitter drive. The transistor then comes out of saturation and rapidly switches off because of the feedback action. The energy in the transformer is dissipated in the damping circuit formed by the diode D and the resistor R in the interval between pulses.

The design of blocking oscillators is discussed by Linvill and Mattson.[96] A number of pulse-generator circuits based on blocking oscillators, with pulse rise times as fast as 0·1 microsec at repetition frequencies up to 50 000 pulses/sec and pulse widths variable from 0·4 to 9 microsec, are given by Eckess.[97] Techniques for the generation of sawtooth waveforms have been described by Stephenson[98] and others.[99]

High-speed switching circuits have also been achieved by the use of *p-n-p* and *n-p-n* transistors in a hook connection (corresponding to *p-n-p-n* devices).[3] A compound circuit of this form has an overall current gain, α, greater than unity and may be

used in switching circuits originally based on point-contact units. One possible circuit arrangement is described by Henle.[100]

(7.3) Transistor Counting Circuits

The development of scaling circuits using transistors is a field of appreciable general interest.

Krenitsky has described circuits operating neon indicator lamps in ten- or four-lamp decade scalars using silicon transistors.[101] The binary stages are conventional flip-flop arrangements with diode clamps and they require input pulses of 10 volts. The circuit was satisfactory up to 80° C for counts in the range 5–200 kc/s.

More recently Chaplin and Owens described a basic binary scaler that uses transformer coupling, thus avoiding the limitation on speed imposed by capacitance coupling.[102] With currently available low-frequency junction transistors ($f_\alpha \simeq 500$ kc/s) the circuit is capable of reliably accepting pulses ranging in width from 0·3 microsec to over 1 microsec at a minimum amplitude of 1·5 volts, the resolution being 2 microsec. A complete scale-of-ten circuit uses eight transistors, ten diodes and five transformers.

(8) TRANSISTOR CONTROL OF MAGNETIC AMPLIFIERS

Some interesting possibilities exist for the incorporation of transistors in magnetic-amplifier circuits.[103, 104, 107] In Fig. 16(*a*)

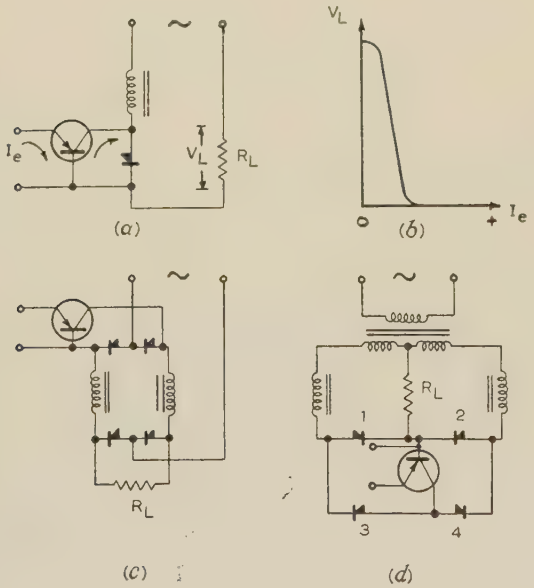


Fig. 16.—Transistor control of magnetic amplifiers.
(*a*) Control of auto-self-excited magnetic-amplifier element.
(*b*) Resulting control characteristic.
(*c*) Bridge-connected magnetic amplifier with transistor control.
(*d*) Centre-tap magnetic amplifier with transistor control.

a single-core magnetic amplifier is shown with the auto-self-excitation rectifier shunted by a transistor. Variation of emitter current controls the mean output voltage of the magnetic amplifier as shown in Fig. 16(*b*). Bridge-connected and centre-tap magnetic amplifiers similarly controlled are shown in Figs. 16(*c*) and 16(*d*). The transistor, which may also be applied in the common-emitter configuration, supplies only the magnetizing current for core flux changes. In certain applications a germanium-junction photo-electric cell may be used in a similar way to give a simple but effective circuit.[105]

In reset (Ramey) magnetic amplifiers, control of the output

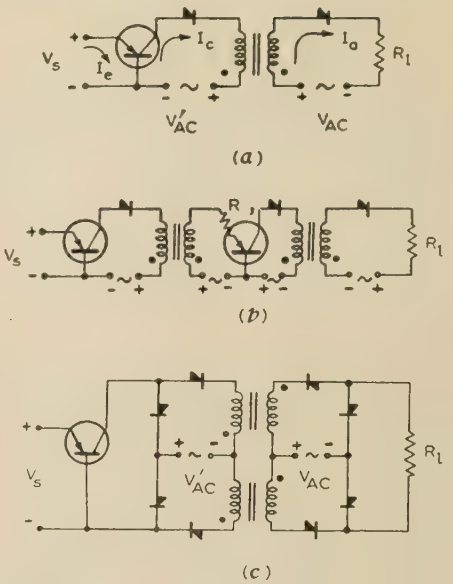


Fig. 17.—Transistor control of reset magnetic amplifiers.
(*a*) Single-core circuit.
(*b*) Coupling between single-core stages.
(*c*) Control of full-wave circuit.

by change of resistance in the control circuit is well known, and Fig. 17(*a*) shows the application of this principle using a transistor. If the emitter current is zero the collector current is also virtually zero and no flux reset action takes place. Full output of the magnetic amplifier is thus obtained. The voltage across the transistor during the reset half-cycle is therefore the reset supply voltage V'_{AC}, and the transistor may be regarded as having maximum impedance. When emitter current is applied, collector current flows, permitting some reset of the core flux, and therefore the output during the load half-cycle is decreased. The voltage across the control winding depends on the flux-resetting process, and the voltage across the transistor collector-base terminals is the difference between this voltage and the reset supply voltage V'_{AC}.

Coupling of two single-core circuits in cascade may be arranged as shown[106] in Fig. 17(*b*). If a single-stage full-wave output is required the circuit in Fig. 17(*c*) may be used—this is a conventional two-core reset magnetic amplifier with transistor control of the reset current. In magnetic-amplifier single-stage circuits power gains of over 10^3 are normal for units handling tens or hundreds of watts, and therefore readily available transistors with dissipation ratings up to a few hundred milliwatts are suitable for controlling magnetic amplifiers of considerable power ratings.

In servo-amplifiers operating from a.c. input signals the present design trend is to use transistor pre-amplifiers followed by phase-sensitive rectification to convert the signal to direct current which is then applied to the control windings of the magnetic-amplifier output stage.[37] The conversion to direct current provides the opportunity for inserting phase-shaping networks for servo-stabilization before the signal is passed on to the magnetic amplifier. However, the servo-amplifier described recently by Weiss is an interesting example of a transistor pre-amplifier controlling with an alternating-voltage output a magnetic-amplifier stage to which phase-advance and integral control are applied.[108]

The retention of magnetic-amplifier output stages in servo-amplifiers is principally indicated where the power requirements

are in excess of those obtainable from readily available power transistors or where the ambient temperature conditions are severe. The provision of transistor pre-amplifiers in place of magnetic-amplifier input stages is normally made with the object of improving servo-mechanism characteristics and reducing amplifier size.

(9) SWITCHED TRANSISTOR CIRCUITS FOR POWER CONTROL

Junction transistors with suitable base drive may be made to conduct fully with low voltage-drops (a few millivolts to say half a volt, depending on the device and the type of application), and with appropriate reversal of base bias they can be transformed rapidly to almost perfect cut-off. The possibilities of controlling power by such switched action were expounded by Alexanderson[109] in 1952, and somewhat later Ebers and Moll presented a basic theory of large-signal behaviour of transistors.[59] Appreciable power-handling ability is retained up to high ambient temperatures with switched operation because the power loss in the transistor is low. Furthermore, changes of parameters such as α and I_{c0} with temperature, etc., have limited influence on circuit action in this mode of operation.

Apart from the low-power-level switching circuits described by Bright (Section 6), the principal power-switching uses of transistors have been in square-wave saturable-transformer-type oscillators (Section 4.1). More recently, amplifier and regulating circuits involving switched operation have been devised.

(9.1) Amplifiers with Switched Action

Initial work on a class of transistor amplifier with switched mode of operation, the switching angles being determined by the combination of sinusoidal signal and bias voltages, has been described elsewhere.[110] Many of these amplifier arrangements are similar to magnetic amplifiers in circuit configuration and output waveforms. Certain of the transistor circuits examined, however, are not directly related to magnetic-amplifier circuits, since the latter are limited by core flux reset requirements whereas transistor configurations are not. More recently, switching principles have been applied to class-B push-pull transistor-amplifier configurations; and since these provide phase-reversing a.c. output characteristics, the new arrangements are of possible interest in servo-motor control.

Fig. 18 shows the circuit arrangement, operating waveforms and theoretical characteristics of a class-B configuration with switched action. In Fig. 18(*a*) the difference from a normal class-B circuit is that bias voltage of the waveform in Fig. 18(*b*) is provided in the base drive circuit. This ensures transistor cut-off until the sum of the instantaneous signal and bias voltages is positive, as at θ in Fig. 18(*d*). Since the signal and bias voltages are chosen to be of appreciable amplitude relative to the base drive voltage required for saturation (say, 5 volts relative to 0·4 volt), the output voltage rises rapidly to the value of the supply voltage within a short period after the onset of switching at θ. The output continues at this value, as shown in Fig. 18(*d*), until the instantaneous sum of signal and bias voltage decreases to zero and becomes negative at $\pi - \theta$, when cut-off occurs. The switched output waveform is in phase with the signal voltage and reverses with reversal of the signal phase.

The equation determining θ is

$$V_s \sin\theta + V_b(\sin\theta - 1) = 0 \quad . \quad . \quad . \quad (13)$$

and the amplitude of the fundamental components of output voltage (assuming complete switching at θ and $\pi - \theta$) may be calculated by Fourier analysis to derive the theoretical characteristics shown in Fig. 18(*e*). The full-line curve in Fig. 18(*e*) is

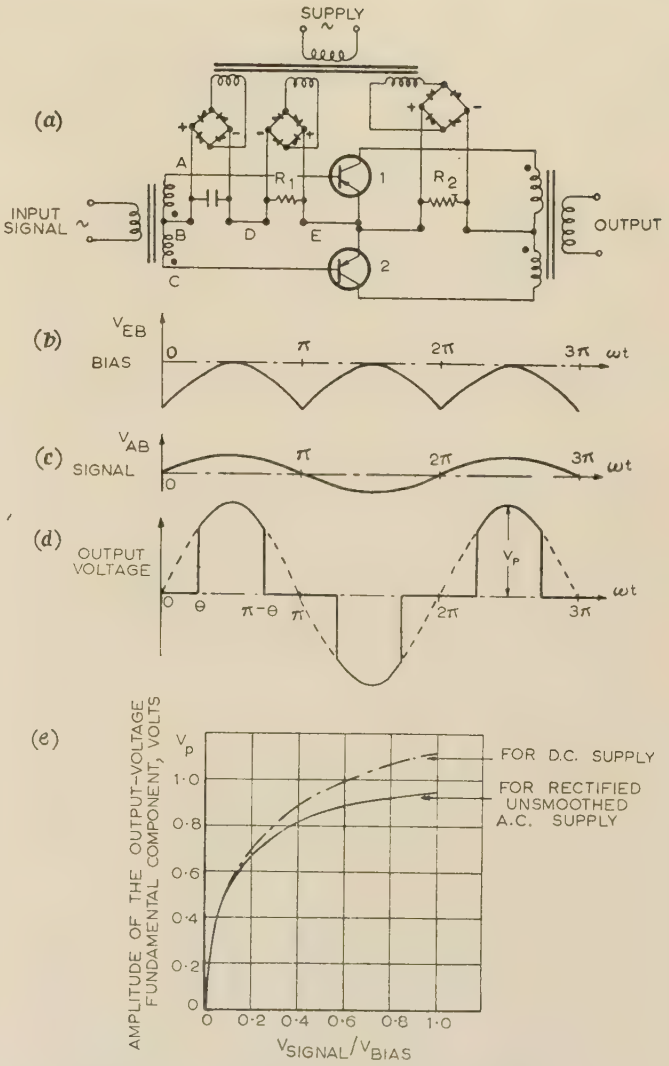


Fig. 18.—Switched class-B transistor amplifier.
Bias arranged to give high-gain characteristic.
(*a*) Circuit arrangement.
(*b*) Bias voltage in the base-drive circuit.
(*c*) Signal voltage.
(*d*) Resulting output voltage waveform.
(*e*) Theoretical output characteristics.

for a rectified unsmoothed a.c. power supply, and the chain-dotted line is for a smoothed d.c. supply.

A second version of the switched class-B amplifier has been devised which has a more linear characteristic and lower gain. The circuit arrangement is given in Fig. 19(*a*) and the bias waveform is that of Fig. 19(*b*). If the capacitance C_1 provides nearly 90° phase shift, the switching angle θ is determined by the condition

$$V_s \sin\theta - V_b \cos\theta = 0 \quad . \quad . \quad . \quad . \quad (14)$$

and the theoretical output characteristic is that given in Fig. 19(*e*) with a load waveform as in Fig. 19(*d*).

For this latter circuit, typical operating magnitudes may be an output voltage of 15 volts (r.m.s.) into 10 ohms impedance for a signal of 5 volts (r.m.s.) at 50 ohms impedance—this represents a power gain of 16 dB. The circuits are still in a development stage but show promise as output stages for transistor servo-amplifiers.

An alternative line of development lies in pulse-width modu-

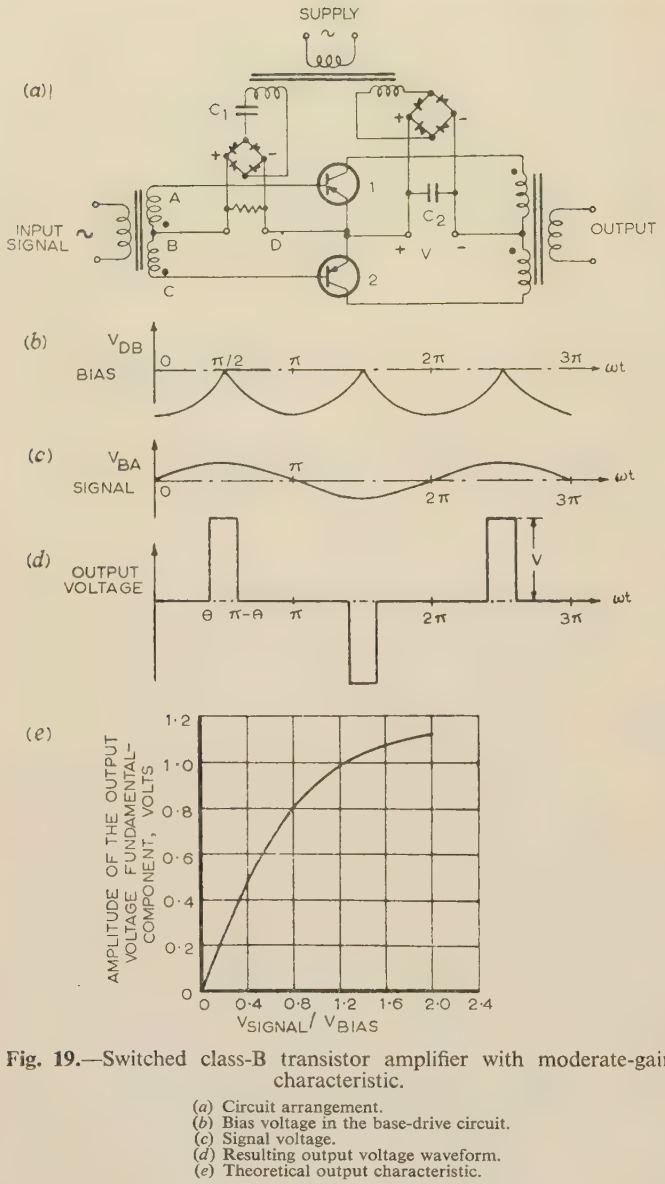


Fig. 19.—Switched class-B transistor amplifier with moderate-gain characteristic.

(*a*) Circuit arrangement.
(*b*) Bias voltage in the base-drive circuit.
(*c*) Signal voltage.
(*d*) Resulting output voltage waveform.
(*e*) Theoretical output characteristic.

lation of amplifier power transistors by the use of magnetic amplifiers, preferably operating from high-frequency square-wave supply sources.[111] The magnetic amplifier, in addition to its pulse-width modulation feature, provides considerable pre-amplification; and at the supply frequencies normally used, say 1–10 kc/s, the amplifier and its power supply are conveniently small in size and weight. The maximum practical switching rate is determined by transistor switching losses rather than by the magnetic amplifier. Magnetic amplifiers may be arranged to accept control by either d.c. or a.c. signal inputs and have zero stabilities of between 10^{-6} and 10^{-9} watt input power, depending on the circuit configuration.

(9.1) Power-Supply Regulators with Switched Action

The power-regulating systems described in Section 3.1 operate in linear fashion, and considerable power losses tend to occur in the regulating transistors. Switched operation of power supply regulators is therefore attractive.[112]

Fig. 20 shows the circuit arrangement of a switching regulator designed by the author to give an output of 20 volts ±1% for

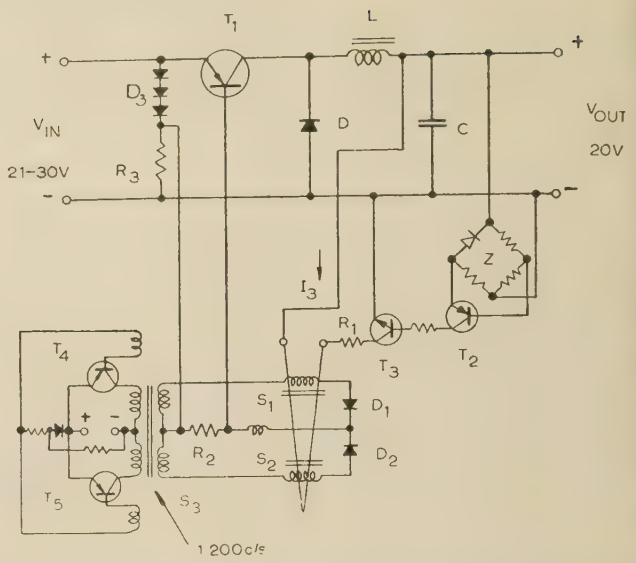


Fig. 20.—Voltage-regulated supply with switched transistor operation controlled by a magnetic amplifier and reference-diode bridge.

an input voltage range of 21–30 volts with a load current range of 0·2–2 amp. The output voltage is sensed by the silicon reference-diode bridge Z, which causes the collector currents of T_2 and T_3 to rise if the voltage is in excess of the chosen regulated value. The current I_3 changes by 5 mA for a change in the output-voltage level of 0·1 volt, and this variation is applied to the control winding of a centre-tap magnetic amplifier (S_1, S_2, D_1, D_2) and is sufficient to cause considerable pulse-width modulation of the voltage across the load resistor R_2. The magnetic-amplifier operating frequency is 1 200 c/s, and this is provided by a square-wave saturable-transformer oscillator (T_4, T_5, S_3) operating from the unregulated input voltage. The voltage across the resistor R_2 is unidirectional, of variable pulse width and of 8 volts amplitude. The potential divider chain D_3, R_3 provides a drive voltage for the power transistor T_1 of approximately 3 volts, which causes full conduction in the absence of voltage across R_2. Since the voltage across R_2 is derived from a 1 200 c/s a.c. supply with rectification, and arranged to make the base of T_1 positive with respect to the emitter, the effect is that non-conducting periods of T_1 are produced at 2 400 pulses per sec with a pulse length that regulates the mean output to 20 volts. Smoothing is accomplished by the inductor L, the capacitor C (50 μF) and the diode D. The overall circuit efficiency is 70% at full load, and the complete unit measures 7 in × 4 in × 3 in and weighs 2¾ lb.

If such regulated voltage supplies are applied to square-wave saturable-transformer oscillators, constant frequency and voltage outputs are obtained. Similarly the technique may be applied to d.c.-to-d.c. convertors to give regulation of the h.t. output voltage.

(10) CONCLUSIONS

Techniques of transistor circuit analysis based on equivalent circuit networks, and the fundamental principles of transistor circuit design and operation, are well established.[2, 3]

In low-power amplifier circuits, techniques of overall and local feedback to give gain stability with temperature, and desired shaping of gain/frequency characteristics, have been developed. Automatic gain control is available when required, and zero drift correction may be used in operational amplifiers. In the servo-mechanism field, amplifiers of 5 watts (or more) output power are under development for 2-phase motor control.

Transistor pre-amplifiers with magnetic-amplifier output stages are also used.

Increased power ratings have permitted the use of transistors in voltage and current regulators, sometimes in conjunction with magnetic amplifiers, and silicon avalanche diodes may be used as reference elements.

In feedback oscillators the circuit trends have been in the direction of securing amplitude and frequency stability (up to one part in 10^5) and low harmonic distortion when required. Three-phase oscillator circuits capable of substantial power outputs have been developed.

An important development has been appreciation of the nearly ideal switching characteristics of germanium-alloy junction transistors. Chopping circuits suitable for millivolt signal levels, and high-performance demodulators, have been devised. Such circuits operate satisfactorily over wide temperature ranges, say $-65°$ C to $+70°$ C.

Switched operation implies high efficiency and low dissipation loss, and numerous power-switching circuits have been devised. Saturable-transformer square-wave oscillator circuits for d.c.-to-a.c., or d.c.-to-d.c., power conversion are important examples and other circuits for amplifiers and regulators are described.

The switching properties of transistors are being applied in conjunction with magnetic cores for various analogue and digital computing purposes. Incorporation of transistors in digital computers is bringing about considerable size reduction, the power-supply requirements being less than for thermionic-valve circuits by a factor of a thousand or more.

Although transistors with acceptable characteristics are now available, for many applications overall improvements are still required in ratings, particularly in respect of cut-off frequency. Silicon transistors are needed in many military applications to meet Service temperature requirements. In the entertainment and industrial fields one of the major problems facing manufacturers is that of reduction of cost.

(11) ACKNOWLEDGMENTS

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DISCUSSION BEFORE THE RADIO AND TELECOMMUNICATION SECTION, 15TH MAY, 1957

Dr. E. V. D. Glazier: I have been interested in the application of the transistor to military equipment, since the first announcement of the transistor effect in America. My interest has related to two broad categories—the pulse type of circuit using digital techniques and the class-A amplifying technique.

In the early days we tried, with both British and American point-contact transistors, to design shift-register and counter circuits. They had to operate certainly up to 55° C ambient temperature, and preferably up to 60 or 70° C, and over reasonable battery voltage variations.

After about three years shift-register circuits with point-contact transistors had been developed to meet these requirements, but the power consumption was about 0·5 watt per stage. Using thermionic valves, power consumptions were about 3 watts per stage, and so this represented an advance, but we felt the need for something better.

About 1954, when it was still being said that point-contact transistors should be used for digital techniques and junction transistors for class-A amplifying techniques, we decided to use junction transistors for digital circuits. Immediately designers could meet the temperature and voltage scale, the tolerancing was not too difficult, and the power consumption was down to 20 mW per stage.

We realized that it was necessary to do life tests, and so we built a circuit consisting of 42 transistors, in a shift-register arrangement, so that even one misoperation would be recorded, and we just let it run. It has now done some 20 000 hours without a transistor fault or, for that matter, any other kind of fault. Using the author's figure of 1% per 1 000 hours, the probability that we should have done so well is about one part in 20 000. This is fairly good evidence that the figure of 1% in 1 000 hours is rather pessimistic. If we take a figure of 0·1% per 1 000 hours, a probability of 0·4 is obtained, which perhaps means we should get a fault before long.

The experiment was carried out at room temperature. We should have set up a parallel test running at 55° C all the time. I believe this has been done, and I would like to have the result.

We were so impressed with the Eccles–Jordan circuit that we committed ourselves to its use. With this circuit we could tackle equipments that would have contained several hundred thermionic valves, but using valves the reliability would always have been a doubtful factor. With transistors, however, these equipments could be achieved.

It is true to say that we have really had no regrets. We are having second thoughts about core and transistor combinations, particularly the type of shift register shown in Fig. 13. It is more economical in components than Eccles–Jordan circuits, but probably these very small ferrite cores are still not easy to obtain. However, I believe that all circuits consisting of transistors and ferrite cores in combination involve a fairly heavy current pulse through the transistor, and I wonder whether we should get such good transistor lives.

I agree with the author that, for digital and data-handling techniques, the transistor has already replaced the thermionic valve.

I should have liked more details in the paper on high-grade amplifier applications, i.e. the type of amplifiers that might be used for radio or radar intermediate frequencies.

In America the noise-factor gap between the transistor and the valve is now only 2 or 3 dB for typical 5 or 10 Mc/s i.f. amplifiers. I wonder whether this is a fundamental gap, or can we really look forward to transistors with noise figures comparable to those of valves?

Even if we are a year or two behind America in the production of transistors, particularly the higher-frequency types, we have been very satisfied with the British product, particularly from the point of view of life and uniformity. It is said that they can be a good deal more uniform than valves.

In America they are looking forward to an expansion of production of something like a scale of two per annum, and if British manufacturers can look forward to the same rate of expansion, they should get a very good financial return.

Dr. G. B. B. Chaplin: In the Summary the author states that 'transistors are capable of supplanting thermionic circuits in all except high-frequency applications (above 10 Mc/s with currently available units)'. I feel that the main limitation of the transistor is its relatively low input resistance, the effect of which is more serious at low rather than high frequencies. For example, the germanium transistor is unable to amplify direct currents of much less than $10^{-2}\,\mu$A, and even the silicon transistor is not expected to reduce this figure below about $10^{-4}\,\mu$A. On the other hand the electrometer valve, which is normally used for this purpose, has an input impedance of hundreds of megohms, and can amplify currents as low as $10^{-8}\,\mu$A—an improvement of several orders of magnitude over the transistor.

At the other end of the frequency spectrum, however, the outlook for the transistor is much more promising. Transistors have already been made in experimental quantities with cut-off frequencies of several hundred megacycles per second. With cut-off frequencies of this magnitude, the limitation to performance is mainly determined by the figure of merit, g_m/C, and it is interesting to compare the transistor and valve on this basis. A typical miniature pentode has mutual conductance, g_m, of up to 10 and an anode output capacitance, C, of 10 pF, resulting in a figure of merit of unity. The transistor has a mutual conductance of approximately $1/r_b$ which is typically 20, and an output capacitance of about 2·5 pF, representing a figure of merit eight times that of the valve.

The transistor therefore appears to have a considerable advantage over the valve in high-frequency circuits, and may well open up applications which have not hitherto been possible using thermionic valves.

Mr. T. R. Scott: The stage is now being reached when the type of transistor has to be taken into account in discussing the circuit design. The paper is obviously written, in the main, round alloyed-junction transistors; the top frequency quoted for available commercial transistors of 10 Mc/s is therefore not surprising. It is unlikely that this will be more than doubled for the alloyed-junction type until the *p-n-i-p* types are available towards the end of the year. For still higher frequencies we must await the commercial production of diffused-base types and the like, say in 1958.

One rather surprising omission is the symmetrical transistor. Reference 4 is not supported by any account in the paper. 'Non-symmetrical units' are referred to in Section 6.1, and so the author presumably admits the existence of symmetrical types. These are particularly useful in switching circuits and in transistor-ferrite arrangements, and were mentioned in the recent Conventions on Ferrites and Digital-Computer Techniques at The Institution. The reference in Section 6.3 to the combination of magnetic cores and transistors is welcomed. This useful combination is too often neglected in surveys.

I also welcome the frank statement that all thermal problems will not be solved by the introduction of the silicon transistor. This is unfortunately true, and every effort must be made to reduce the temperature-variable parameters to suitably low values. The rating of germanium transistors must continue to

be a matter of controversy for some time. First, strenuous efforts are being made to reduce overall thermal gradients below 5° C per watt, with some success. Secondly, the permissible junction temperature quoted by a manufacturer must be based on the lift-test data taken by the manufacturer and on the life which he expects to get. This is true for all components. If a life of the order of 20 years is desired, the rating must be conservative. For some purposes, however, much shorter lives are permissible, and the rating can accordingly be increased.

Mr. A. J. W. M. van Overbeek (*Netherlands*): It seems possible to use transistors in nearly every field of electronics, and, in many cases, they give a considerable improvement. However, in some important respects, transistor circuit design is not yet well established.

One major problem is the noise. It may not be very important at intermediate frequencies, but it is at very low frequencies, as well as near the α cut-off frequency.

Also at high switching frequencies transistors may show memory properties with time-constants which are too long to be attributed to classical minority carrier storage.

On the other hand, there are smaller fields, not mentioned in the paper, in which the transistor has been used in Holland.

One rather promising development is the use of a combination of *p-n-p* and *n-p-n* transistors, the base of the first transistor being connected to the emitter of the second and the collector of the first to the base of the second. A 'hook' transistor can also be used for this purpose. It has the properties of a small thyratron, with the difference that it is much faster. Its trigger voltage is more stable, and, moreover, it can be switched out by a pulse on its control electrode.

Such a thyratron may have a memory lasting only 0·1 microsec. It can be used for pulse-code modulation, for analogue-to-digital conversion and for any other purpose where a small gas triode would be used at lower frequencies.

Another promising circuit comprises two transistors in a cascode circuit at high frequencies, the collector of the first being connected to the emitter of the second. It may give an amplification which is as large as two separate stages, but the circuit is much simpler and can be easily controlled by shunting the input of the second transistor without influencing the input or output circuits of the cascode stage.

Another circuit uses a light-sensitive amplifier or an oscillator. A photo-transistor may be used at the same time as an amplifier or oscillator. As the amplification may be varied more than 80 dB, it is possible to use this property for telecommunication purposes. As an oscillator, it can be used not only for musical instruments but also for translating coded or printed information into a combination of signals with different frequencies.

Mr. A. Heetman (*Netherlands*): The transistor is providing the key to control in certain applications in telephone exchanges. When using a transistor with two stable conditions, e.g. a point-contact or a *p–n–p–n* transistor,* it is possible to switch the transistor from the 'cut-off' to the 'on' condition by applying a high voltage between the emitter and collector.

The principle may be used in a crossbar-switch type of array of transistors (see Fig. A).

When a positive voltage via A and a negative voltage via B are applied, for example, to A_o and B_o the transistor T_{oo} turns over to the 'on' condition.

The other transistors do not 'turn over' because they get only a negative pulse to the collector or a positive pulse to the emitter. Thus no coincidence of negative and positive pulses occurs, and the voltage difference between the collector and emitter is not sufficient to switch to the 'on' condition.

* *Proceedings of the Institute of Radio Engineers*, 1956, **44**, p. 1174. U.S. Patent No. 2.760.004.

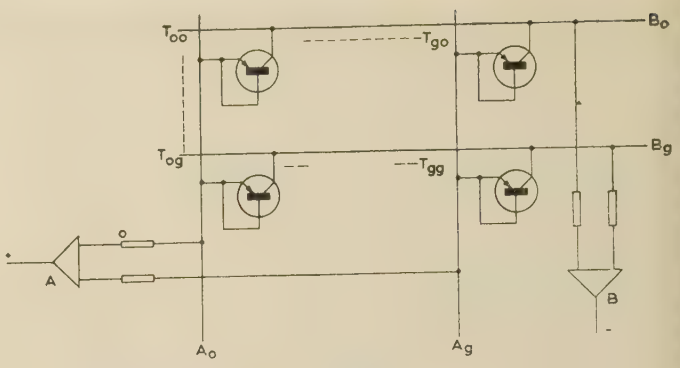


Fig. A

Mr. W. L. Stephenson: Much has been said recently about the desirability of symmetrical transistors for fast switching. The basis of these remarks is the following formula first derived by Moll (Reference 92 of the paper) for the time that a transistor remains in the bottomed state after being switched off at the base:

$$t = \left[\frac{\frac{1}{\omega_n} + \frac{1}{\omega_i}}{1 - \alpha_n\alpha_i}\right]\left[\log_\varepsilon \frac{\alpha_n'(i_{b1} - i_{b2})}{i_c - \alpha_n' i_{b2}}\right]$$

The expression in the first bracket, which may be called the 'hole storage time', depends on the transistor, and the expression in the second bracket depends on the operating conditions. It has been stated in Reference 59 of the paper that, for t to be small, $(1/\omega_n + 1/\omega_i)$ must be small, and the expeditious way of achieving this is to make the transistor symmetrical.

An alternative approach is to say that, for t to be small, $1 - \alpha_n\alpha_i$ should be large, and since from other considerations, particularly that of a fast switch-on, α_n is required to be as near unity as possible, α_i should be as small as possible, i.e. the transistor should be very unsymmetrical.

To take a practical example, a particular type OC45 transistor had the following measured parameters:

$$\alpha_n = 0{\cdot}983$$
$$\alpha_i = 0{\cdot}924$$
$$\omega_n = 3{\cdot}14 \times 10^7$$
$$\omega_i = 1{\cdot}26 \times 10^7$$

This gives a hole storage time of 1·2 microsec.

If this transistor had been symmetrical, with reverse characteristics identical to the forward characteristics, the hole storage time would have been 1·9 microsec, i.e. with a good symmetrical transistor the effect of the increase in ω_i is more than counteracted by the increase in α_i.

Further inspection of the expression for the hole storage time shows that it is more sensitive to variations in α_i than ω_i, so that the unsymmetrical transistor will have the better performance.

It can be seen that the expression in the second bracket can be made very small by making the final base current i_{b2} large and negative. In such a case, reversing the sign of i_{b2}, the expression in the second bracket becomes

$$\log_\varepsilon \frac{\alpha_n'(i_{b1} + i_{b2})}{i_c + \alpha_n' i_{b2}} = \log_\varepsilon \left(1 + \frac{\alpha_n' i_{b1} - i_c}{i_c + \alpha_n' i_{b2}}\right) \simeq \frac{\alpha_n' i_{b1} - i_c}{i_c + \alpha_n' i_{b2}} < \frac{i_{b1}}{i_{b2}}$$

so that with sufficient reverse base drive on switching off, the bottoming delay can be reduced below the transistor hole storage time by a very large factor. There is no difficulty, in

practice, in making i_{b2} large, since the only limiting factor is the ohmic base resistance. Since this is usually of the order of 100 ohms, an applied reverse bias of only 1 volt can cause a reverse base current of about 10 mA.

Mr. J. N. Barry: Whilst certain applications are covered very adequately in the paper, there is very little discussion of the entertainment field, although up to now this has represented quite a major application of transistors. Gramophone amplifiers, radio receivers, etc., are hardly mentioned, and television applications are not mentioned at all.

Recent advances in the power and frequency operation of transistors indicate that television applications are beginning to be possible, and Sziklai, Lohman and Herzog* described an all-transistor portable television receiver as long ago as 1953.

There are references to work on radio receivers, but most refer to circuits with eight or nine transistors per receiver, and I do not think they can be considered typical at present. A typical battery portable receiver at present would probably incorporate a total of six transistors. There is more recent information to be found in the German literature.†

Cost is mentioned as a limiting factor in the entertainment field, though there is little evidence given to substantiate this statement. Part of the reason is not only the high cost of the transistor, but also the special circuits that are needed to overcome some of the disadvantages of transistors, such as variations of the parameters with temperature and working voltage.

There are a few generalized statements in the paper which are open to some doubt at present. These include comments on transistors 'supplanting thermionic circuits in all except high-frequency applications (above 10 Mc/s with currently available units)', the rise in collector-voltage ratings (in most cases there has been a tendency towards a reduction in rating with higher-frequency operation) and a comparison between the power consumptions of transistor and valve receivers.

It is a pity that no indications have been attempted about future possibilities as a whole. Has the author any comments to make on the relative future importance of transistors in different fields, e.g. the entertainment industry, the telephone industry and military applications?

Mr. L. P. Morgan: I disagree slightly with one of the author's conclusions, in which he states that techniques of transistor circuit analysis based on equivalent circuit networks, and the fundamental principles of transistor circuit design and operation, are well established. The most obvious evidence that this is not necessarily so is borne out by the number of different parameters that are considered useful. Some consider the transistor as a 'black box' and use either the *h*-, Y- or Z-parameters. Others prefer to use a circuit which represents the physical action of the transistor, and then the T- or the hybrid-π circuit must be considered. This indicates that there is still much work to be done in deciding which of these parameters and circuits are the most useful.

Much work is still to be done on the fundamental equivalent circuits of the transistor. It can safely be assumed that they are not correct, since there are discrepancies of some 10–20% between the measured Y-parameters and the values predicted from the equivalent circuits.

Another important fact is that the transistor cannot be considered as just another valve. Some people treat the valve as a device whose method of operation need not be fully understood, and which is, in fact, a box giving an output current proportional to the input voltage. This is probably an over-statement, but the transistor circuit engineer cannot take this attitude, and really must know quite a lot about the workings of the transistor to realize its full potentialities.

Dr. A. J. Biggs: Some conflicting views have emerged on the frequency ranges in which transistors will fit best for applications of the future. So far as the entertainment field is concerned, the suitability of transistors cannot be assessed solely on a frequency basis. For example, it is likely that, for many years to come, the line output stage of a television receiver operating at 10 kc/s and the radio-frequency amplifier operating at 200 Mc/s will be more satisfactorily carried out by thermionic valves. The electronic engineer must therefore accept the transistor as a new device for certain applications but not as something which will entirely supplant the thermionic valve.

Dr. F. L. C. White: I should like to put in a plea for the manufacture in this country of at least a few types of *n–p–n* transistors. The ability to choose the polarity at will can often lead to simplification in circuit design, as shown, for example, in Figs. 3(*a*) and 3(*b*).

It seems preferable to continue to use the circuit convention of drawing the more positive parts above the more negative parts. To reverse the usual order, as in Figs. 6(*a*), 8(*b*), 9 and 15, simply because *p–n–p* transistors are used, can be very misleading when trying to interpret waveforms in relation to the circuit, and will lead to difficulty when *n–p–n* transistors or thermionic valves are used with *p–n–p* transistors in one circuit.

Mr. A. R. Owens: With reference to Dr. White's remarks, the use of a *n–p–n* transistor may be rendered unnecessary in a direct-coupled circuit in certain cases.

Examining the collector characteristics of a typical *p–n–p* a.f. transistor (Fig. B), it may be seen that current gain, as evidenced

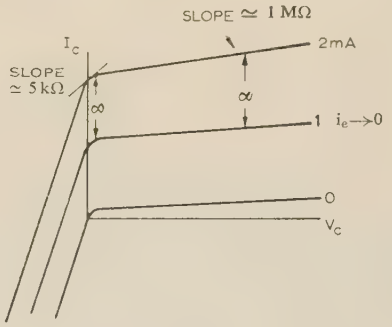


Fig. B

by the spacing between the lines of constant emitter current, is maintained at collector voltages down to zero. There is thus no need to apply a collector–base potential to obtain current gain.

The collector impedance r_c is given by the slope of the collector characteristic, and even at $v_c = 0$, this impedance is of the order of several kilohms.

The circuit of Fig. C is therefore quite feasible. The d.c. negative-feedback path, R_1 and R_2, ensures that the circuit operates in the linear region. Since the emitter–base voltages are approximately equal for all three transistors, the collector–base voltages of J_1 and J_2 must be zero, and as described above, current gain is still available.

The transfer of current from one stage to the next depends upon the ratio of its output impedance to the input impedance of the following stage. The output impedance is lowered by reducing the collector voltage, and the current gain may be considerably reduced.

However, if transistors with lower leakage currents are used,

* Sziklai, G. C., Lohman, R. D., and Hertzog, G. B.: 'A Study of Transistor Circuits for Television', *Proceedings of the Institute of Radio Engineers*, 1953, **41**, p. 708.
† Olschewski, R.: 'Kofferempfänger mit Transistoren', *Electrotechnische Zeitschrift*, **78A**, 1957, p. 20.

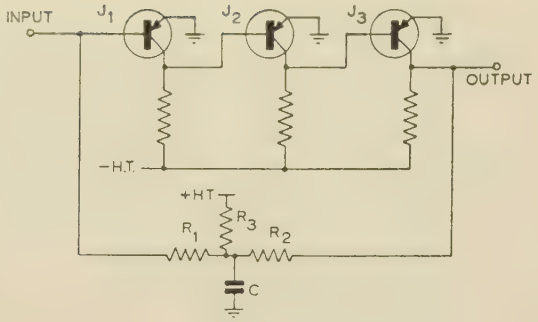


Fig. C

such as high-frequency units, the output impedances are correspondingly higher, and the full gain may be obtained.

In Fig. C the capacitor C eliminates a.c. negative feedback, and the current flowing in R_3 ensures that the collector of J_3 is biased negatively in the normal manner, thus allowing a considerable output voltage swing.

With regard to the complementary emitter-follower, in the case where it is being driven from a previous transistor stage, the *n–p–n* transistor may be replaced by a diode (Fig. D).

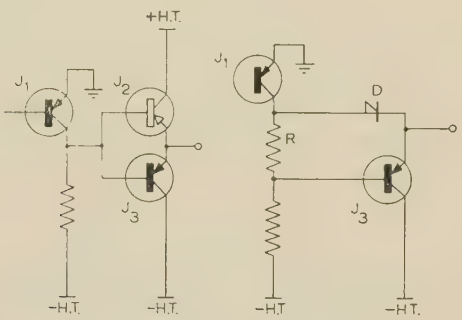


Fig. D

The emitter follower J_3 provides current for negative-going pulses as before, but the drive stage J_1 now provides current for positive pulses through D. A resistor R may be included to provide bias which ensures that both D and J_3 are not cut off simultaneously.

Mr. D. E. A. Harvey (*communicated*): It is to be regretted that the paper shows such a bias towards non-entertainment aspects.

The advantages of semi-conductor devices in entertainment applications are perhaps not quite so positive as in certain industrial and military fields, and the applications engineer is presented with considerable problems in his efforts to produce transistorized equipment which is competitive, in the eyes of the public, with that employing thermionic valves.

An instance occurs in the case of the portable transistorized broadcast receiver which is now appearing on the British market. Early American versions tended to sell merely on novelty value, and there was little incentive for manufacturers to streamline their designs. As many as eight or nine transistors were often used in each receiver, and yet reproduction was usually poor by modern standards.

Considerable research has now been devoted to the design of such receivers, and the latest British designs are in no way inferior to their thermionic counterparts. The transistor complement has been reduced to five or six, and running costs are so low relative to the quality and volume of a.f. output that the receiver can be considered as a competitor to the mains/battery type of transportable receiver, which makes the initial cost more attractive than it would otherwise be.

To reduce costs, self-oscillating frequency-mixer stages have been developed which give a conversion gain only 2 dB less than a mixer-plus-oscillator combination, while the noise performance compares more than favourably with that of a battery-operated pentagrid convertor. Amplified a.g.c. circuits have been devised which make use of one of the later stages to amplify the d.c. component of the rectified signal without interfering with the fundamental operation of the stage. Very good a.g.c. characteristics result.

A.G.C. arrangements which employ collector-voltage control are not favoured for entertainment applications, because of the relationship between collector junction capacitance and junction voltage, $C_c \propto 1/\sqrt{v}$. This leads to changes in neutralizing conditions which may result in distortion of response shape or even complete instability.

Mr. M. G. Scroggie (*communicated*): The transistor symbol used in the paper* is very suitable for point-contact transistors. It is, however, most unsuitable for junction transistors, first, because the use of one symbol for two very different kinds of component necessitates study of the context to determine which, in any particular occurrence, it is intended to mean; and secondly, because its appearance in no way suggests what it represents. On the contrary it strongly suggests something it is not. Moreover, it is not readily modified to represent the many special types of junction transistor coming into existence.

The use of this symbol for junction transistors is equivalent to using the term 'point-contact transistor' to mean also 'junction transistor', and relying on the context to make the distinction. Even if point-contact transistors were regarded as obsolete, such a practice would be absurd.

The objection that the practice is well established has always been raised against any sort of progress: the slave trade was once well established in the British economy. In our own time and field, minor anomalies such as use of the word 'resistance' to refer to a component have readily been corrected, and at a later stage in the art than that now reached with transistors. Several convenient and suggestive systems of junction-transistor symbolism have been devised; all that is needed is for the standardizing committees to settle on one.

Mr. K. P. P. Nambiar (*communicated*): The criticism of the speaker who referred to the absence of a transistorized line timebase for scanning television receiver tubes appears to be very harsh indeed. Line-scanning circuits have been designed and operated in miniature television cameras† as well as in wide-angle receivers.‡ In fact, a circuit§ which was found to be adequate for scanning a vidicon camera tube has been described in the *Journal*. The transistors used in that particular circuit are the low-power OC72 type. Transistors capable of handling much higher powers are currently available. My own view is that television receiver circuits are not transistorized, owing, not so much to the inability of transistors to perform the appropriate functions, as to the fact that not much work is being done in that direction. Perhaps this is due to a partisan attitude towards the transistor, similar to that shown by the speaker mentioned.

* See B.S. 530 : 1948, Supplement No. 4 (1956), Item 69.

† FLORA, L. E., *et al.*: 'Transistorised Television Cameras Using the Miniature Vidicon', *RCA Review*, 1956, **17**, p. 469.

‡ GUGGI, W. B.: 'CRT Deflection Circuit Has High Efficiency', *Electronics*, April, 1957, p. 172.

§ 'Linear Sweep Circuits', *Journal I.E.E.*, 1957, **3** (New Series), p. 159.

THE AUTHOR'S REPLY TO THE ABOVE DISCUSSION

Dr. A. G. Milnes (*in reply*): The discussion adds considerably to the value of the paper as a review of current thought on transistors, and since the contributions are mainly authoritative comment, my reply will be short.

Dr. Glazier's evidence on reliability in data handling is valuable, and I welcome his opening contribution. The transistor is not suited for very-high-input-impedance applications, as Dr. Chaplin points out. On the high-frequency aspect he should perhaps have mentioned that voltage, current and power dissipations of h.f. units tend to be low. The paper is concerned with circuits for readily available transistors, and these, as Mr. Scott comments, are virtually all of the alloyed-junction type. There has been so much 'forward looking' in the transistor field that, with all respect to Mr. Barry, it seemed wiser to confine the paper to circuits that are with us all, here and now.

Mr. Van Overbeck's and Mr. Heetman's comments on circuits used in the Netherlands are very much appreciated. The statement that switching theory and experiments may not be entirely in accord is important, and incidentally suggests that the relative merits of symmetrical and non-symmetrical constructions, as discussed by Mr. Stephenson, need experimental confirmation. Mr. Morgan's comment that equivalent circuits of alloyed-junction units require some revision to correct discrepancies of 10–20% in parameters is of interest, and I look forward to more detailed information in due course.

Dr. White's plea for *n–p–n* types in addition to *p–n–p*, for reasons of circuit convenience, is strongly supported; although the circuit ingenuity of Mr. Owens may provide for certain needs.

Since the point-contact transistor is obsolete, my view is that there is no serious objection to the transfer of its symbol to present-day alloy and grown-junction units. Four out of five textbooks on the subject, and a high percentage of published papers, use this symbol. Objections by Mr. Scroggie that the physical construction is not represented in the symbol could be levelled, in one respect or another, against all the alternatives that I have seen proposed; and this symbol can be just as readily extended to include tetrodes, etc., as others. Standardization on new symbols, should a general consensus of opinion show it to be desirable, would certainly require negotiation at an international level, to be effective.

Detailed discussion of the problems involved in the application of transistors in the entertainments industry is omitted because of my limited knowledge in this field, and not, of course, because of any intentional bias. Some useful general reading on transistor radio receivers, and on the sections of television systems where transistors can be most effectively used, is to be found in the recent book by Shea.* I also welcome Mr. Nambiar's other references in this field, and recommend the 1957 I.R.E. National Convention Record, Part 3, for further information.

* Shea, R. F.: 'Transistor Circuit Engineering' (Wiley, 1957).

DISCUSSION ON 'TRANSISTOR POWER AMPLIFIERS'* AND 'TRANSISTOR D.C. CONVERTORS'†

WESTERN CENTRE, AT CARDIFF, 11TH MARCH, 1957

Mr. A. N. D. Kerr: Since the papers were first published nearly two years ago and so much progress has taken place in the interim, will the authors describe the latest position, with particular reference to power applications involving motor loads?

Mr. R. Hamer: The use of transistor power amplifiers in small portable a.f. equipment has been admirably demonstrated. Are transistors likely in the near future to replace valves in high-fidelity audio equipment with output powers of tens of watts? Will the authors quote some harmonic distortion figures for output powers of up to, say, 10 watts obtainable with more-linear transistor power amplifiers? Is it possible to achieve better matching of the characteristics of transistors for use in Class-B push-pull amplifiers by fabricating double-junction transistors, e.g. a single *p-n-p-n-p* unit?

Messrs. R. A. Hilbourne and **D. D. Jones** (*in reply*): Most of the transistor structures now being developed or manufactured for power applications have limitations which restrict the supply voltage to a few tens of volts. In order to handle large powers with low distortion it is therefore necessary that the electrical characteristics of the transistors should remain substantially constant over large variations of signal current. Considerable attention has been given to this problem, with the result that transistors having peak current ratings of up to about 10 amp are becoming available. These devices should be satisfactory for the type of application that Mr. Hamer has in mind, so long as sufficient negative feedback is applied to the amplifier in order to obtain adequate frequency response. Class-A push-pull amplifiers handling output powers of 5 watts at a distortion level of less than 1% have been built without difficulty using transistors having characteristics inferior to those of the newer types.

There would appear to be no advantages in making double transistors, but it is easy to see many disadvantages in such a system. Transistors made by substantially the same process do show differences in characteristics, and there is no reason to assume that such differences would vanish by putting two units in the same container.

There is no simple answer to Mr. Kerr's question. Transistors can work into motor loads so long as the power dissipated does not exceed the manufacturers rating. Particular attention must be given to the variation in load impedance which occurs when the motor is starting or stopping.

Mr. L. H. Light and **Mrs. Prudence M. Hooker** (*in reply*): Recent development work has concentrated on transformer-coupled d.c. convertors,‡ which make use of a transformer as

* Hilbourne, R. A., and Jones, D. D.: Paper No. 1861 R, April, 1955 (see 102 B, p. 763).

† Light, L. H., and Hooker, Prudence M.: Paper No. 1862 R, April, 1955 (see 102 B, p. 775).

‡ Light, L. H.: 'Design and Operation of Transistor D.C. Convertors', *Mullard Technical Communications*, 1956, 2, p. 193.
Wolfendale, E.: 'Transistors and their Application' (Heywood, London, 1957), Chapter 10.
Milnes, A. G.: 'Transistor Circuits and Applications' (see page 565).

the voltage step-up device in place of the ringing choke. They are primarily suited to push-pull operation and are capable of handling tens and even hundreds of watts.* Their output characteristic approximates to a constant-voltage one up to a particular output power, but, should the loading exceed this value, oscillations may cease: if they continue (this depends on the circuit arrangement and the nature of the load), the output characteristic will be modified and the transistors may operate unbottomed. There is then a danger of over-dissipation.

The current requirements of a motor are very heavy during starting. It would be uneconomical to design a transformer-coupled convertor capable of supplying this power in its normal mode of operation, but circuits which on overload continue to oscillate with a falling-voltage characteristic are suitable, provided that the transistors are not affected by the high initial dissipation during starting. Alternatively, the starting load may be reduced by one of the conventional starting devices or the inclusion of a thermistor in series with the motor.

When a ringing-choke circuit can be used in the supply (for motors taking up to some 10 watts), the only problem is that of starting oscillations in the presence of the heavy load. Sufficient forward bias on the transistor base will normally ensure starting.

* UCHRIN, G.: 'Transistor Power Convertor Capable of 250 Watts D.C. Output', *Proceedings of the Institute of Radio Engineers*, 1956, **44**, p. 261.

DISCUSSION ON 'A NEW FORM OF HYBRID JUNCTION FOR MICROWAVE FREQUENCIES'*

Mr. Leo Young (*Maryland, U.S.A.: communicated*): The authors' hybrid circuit is an important extension of the binomial coupler. I was fortunate to know of this work several years ago, when it was just beginning, and also the work of Mr. Gouldson which was based on a report by Lippmann.[A]

I have recently analysed couplers with up to eight branches[B] and obtained exact formulae connecting the characteristic impedances, which ensure perfect directivity and any desired coupling ratio at centre frequency. Perfect series junctions and quarter-wave spacings were assumed.

In the 5-branch coupler with uniform main line impedance, the condition for perfect directivity reduces to

$$P = \frac{2(K + KG^2 - G)}{1 + K^2(1 + G^2) - 2GK} \quad . \quad . \quad . \quad . \quad \text{(A)}$$

where G = Normalized impedance of first branch = 0·1 in this case.
K = Normalized impedance of second branch = 0·425 in this case.
P = Normalized impedance of middle branch = 0·650 in this case.

If we take G and K as above, eqn. (A) gives $P = 0{\cdot}601$ for perfect directivity at centre frequency. This corresponds to a branch guide height of 0·240 in instead of 0·260 in.

The condition for equal power division reduces to

$$(1 - G^2)K(2 - PK) + 2G(1 - PK) + P = 2[1 - PK - GK(2 - PK)] \quad \text{(B)}$$

and is independent of directivity.

The best way to start an empirical design in this case would seem to be to choose G, K and P as nearly as possible in the ratio of the binomial coefficients 1, 4 and 6, as was done by the authors, but in the first instance also satisfying eqns. (A) and (B) exactly. Then gradually greater bandwidth can be achieved at the expense of optimum performance at centre frequency.

The same procedure may with the help of the tables be applied to couplers having up to 6 branches and non-uniform main line, and up to 8 branches with uniform main line. The formulae simplify somewhat for the special cases of 3 dB and 0 dB (cross-over) couplers, and these are also given separately.

Messrs. P. D. Lomer and **J. W. Crompton** (*in reply*): The analysis of branch guide directional couplers described by Mr. Young may be compared with that given in a recent monograph by J. W. Crompton.[C] In this paper, it has been shown how de Moivre's theorem may be used in conjunction with correction curves for end effects given by Harrison[D] to simplify the design of multi-element branch-guide directional couplers. As an example, design data have been calculated for the 5-slot 3 dB coupler described by us. The dimensions of the slots derived by this method agree well with those adjusted experimentally to give 3 dB coupling.

The use of the theory to design couplers with more than five slots is not of great practical interest in view of the difficulty in constructing the narrow slots. This problem, which is encountered to some extent in the 5-slot coupler, will become much more exacting for the narrower branch guides of the 8-slot coupler.

* LOMER, P. D., and CROMPTON, J. W.: Paper No. 2356 R, May, 1957 (see **104** B, p. 261).

REFERENCES

(A) LIPPMAN, B. A.: 'Theory of Directional Couplers', Massachusetts Institute of Technology Radiation Laboratory Report No. 860, December, 1945.

(B) YOUNG, L.: 'Branch Guide Directional Couplers', *Proceedings of the National Electronics Conference, Chicago*, 1956, **12**, p. 723. (The expression for Y for the 5-branch coupler in Table 1 contains a term which was incorrectly given as PKH. It should be PKH^2.)

(C) CROMPTON, J. W.: 'A Contribution to the Design of Multi-Element Directional Couplers', *Proceedings I.E.E.*, Monograph No. 230 R, April, 1957 (**104** C, p. 398).

(D) HARRISON, R. J.: 'Design Considerations for Directional Couplers', Massachusetts Institute of Technology Radiation Laboratory Report No. 724, December, 1945.

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THE FREQUENCY-LOCK A.F.C. CIRCUIT

As a Means of Tracking Signals of Changing Frequency in Conditions of Low Signal/Noise Ratio

By R. LEEK, B.Sc.(Eng.), Graduate.

(*The paper was first received* 14*th February, and in revised form* 14*th June*, 1957.)

SUMMARY

A frequency-lock a.f.c. loop was used to lock to and track a weak c.w. signal of changing frequency in the presence of high levels of noise.

The noise-free operation of the loop as a tracking system is discussed qualitatively from first principles, and this is followed by a theoretical analysis.

The tracking performance of the loop is then considered for the case where the signal is in the presence of noise and the concepts of the 'maximum rate of follow' and the 'mean time-to-unlock' are introduced. An examination of the errors resulting from the noise and from changes in the signal frequency leads to a theory of the loop operation for signals in noise.

A range of practical results is presented which provides agreement with this theory within the expected experimental error.

LIST OF SYMBOLS

$-A$ = Zero-frequency gain of amplifier in shaping network.
α = Angle between locus of maxima and $1/T$ axis on the $(\hat{F}_s, 1/T)$ plot.
$b = \hat{F}_{sv}$.
B_a = Bandwidth between 3dB points of the amplitude characteristic of narrow-band amplifier N.
B_c = Bandwidth between 3dB points of the amplitude characteristic of the discriminator tuned circuit C.
B_n = Noise bandwidth.
c_1 = Change in v_2 per unit difference between the discriminator input signal frequency and the discriminator tuned frequency with $v_1 = 1$ volt (r.m.s.).
C_1 = Capacitance of shaping network capacitor.
$C_{1c} = C_1$ when $T = T_c$.
$\Delta f = f_g - f_s$.
Δf_1 = Steady-state value of Δf following a change in f_s.
Δf_2 = Steady-state value of Δf following a change in F_s.
Δf_{2a} = Directly amplified part of Δf_2.
Δf_{2b} = Integrated part of Δf_2.
$\Delta \hat{f}_2$ = Maximum value of Δf_2.
$\Delta \hat{f}_{2r} = \Delta \hat{f}_2$ when $T = rT_c$.
Δf_n = R.M.S. noise error.
$\Delta f_{nr} = \Delta f_n$ when $T = rT_c$.
$\Delta f_t = \Delta f_2 + \Delta f_n$.
$f_i + f_g$ = Frequency of controlled oscillator output signal.
f_{g0} = 'Free-running' value of f_g corresponding to the standing voltage V_3 at the input to the oscillator control with $v_3 = 0$.
f_i = Centre frequency of narrow-band amplifier N.
f_s = Frequency of input signal to balanced mixer.
f_{s0} = Mean value of a range of values of f_s.
$F_s(p)$ = Operational form of input signal frequency change.
F_g = Rate of change of f_g.
F_s = Rate of change of f_s.
$\hat{F}_s$ = Maximum value of F_s.
$\hat{F}_{sv} = \hat{F}_s$ at a vertex.
$\hat{F}'_{sv}$ = Particular value of $\hat{F}_{sv}$.
$k_1 = c_1 v_1$.
k_2 = Change in f_g per unit change in v_3.
$k_3 = k_1 k_2$.
p = Laplace operator.
$r = T/T_c$.
R_1 = Resistance of shaping network resistor.
R_L = Resistance of resistor in series with v_L.
S = Signal/noise power ratio.
S' = Particular value of S.
S_v = Value of S at a vertex.
t = Time.
t_0 = Time at which f_g had the value f_{g0}.
t_s = Time at which a signal of frequency f_s was applied.
t_u = Time-to-unlock.
$T = C_1 R_1 / k_3$.
T_c = Critical value of T.
T'_c = Particular value of T_c.
v_1 = Input signal to discriminator.
v_2 = Output voltage from discriminator.
v_3 = Output voltage from shaping network.
v_L = External voltage source connected to junction of R_1 and C_1 via R_L.
V_3 = Standing value of the voltage at the input to the oscillator control with $v_3 = 0$.
$x = 1/T$.
$y = \hat{F}_s$.
$Y_s(p)$ = Transfer function of the shaping network, i.e. v_3/v_2.

(1) INTRODUCTION

A requirement which arose in a certain type of control system was the ability to track a signal of varying frequency, in high levels of noise, and a frequency-lock a.f.c. circuit was successfully used for this purpose.

Frequency-lock a.f.c. systems are commonly used in radio receivers where the problem is usually one of correcting the frequency of an oscillator by a fixed amount arising from station switching or drift.[1–11]

The present application is concerned with tracking a higher-order property of the input signal, namely its rate of change of frequency. The magnitude of this frequency ramp was of the order of 5kc/s per second. The results of previous work on this type of problem have been reported by the author concerning the use of an a.f.c. loop of a phase-lock rather than the frequency-lock type.[12]

(2) FREQUENCY-LOCK LOOP

A block schematic of a frequency-lock loop is shown in Fig. 1. The input signal is received at a frequency f_s, and, for simplicity, the loop is first considered to be locked to this frequency. An oscillator O, whose frequency is to be controlled to the locked condition, then operates at a frequency $(f_i + f_s)$. The input signal is mixed with the controlled oscillator signal in a balanced

Written contributions on papers published without being read at meetings are invited for consideration with a view to publication.
Mr. Leek is with the English Electric Co., Ltd.

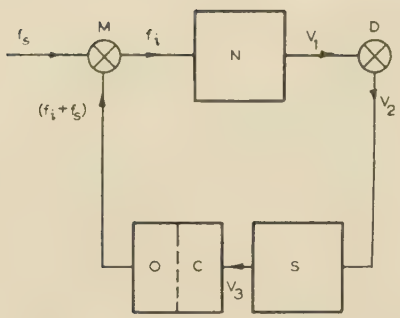


Fig. 1.—A frequency-lock loop.

M : Balanced mixer.
N : Narrowband amplifier.
D : Frequency discriminator.
S : Shaping network.
C : Oscillator control.
O : Controlled oscillator.

mixer M to produce an output at frequency f_i. The mixer output is passed to a band-pass amplifier N whose centre frequency is f_i, and forms the input signal v_1 to a frequency discriminator D. The discriminator is tuned to give zero d.c. output when the input signal frequency is f_i. The input signal v_1 will, in general, differ in frequency from f_i, and the output signal v_2 from the discriminator is proportional to this frequency difference in magnitude and sense. The discriminator is the error-sensing device in the loop, and v_2 is the error voltage which actuates the feedback process. v_2 is passed to a shaping or equalizing network S, and in the simple case this is a single-stage integrator. The output voltage v_3 from the shaping network is applied to an oscillator control C, by means of which the controlled oscillator frequency is made proportional to the magnitude of v_3. v_3 is just sufficient to control the oscillator frequency to $(f_i + f_s)$, and the loop is thus in a stable locked condition.

The loop is effectively a position-control servo mechanism operating on a direct error voltage from the discriminator. Many equalizing networks may be used in such loops, and the

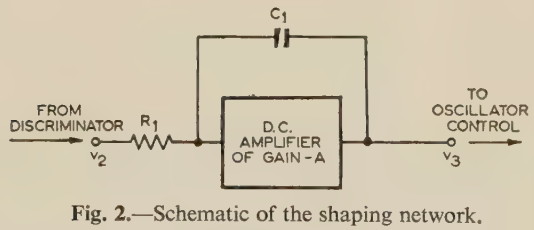


Fig. 2.—Schematic of the shaping network.

Typical values:
A = 130.
$C_1 = 0{\cdot}3\,\mu F.$
$R_1 = 1{\cdot}0\,M\Omega.$

elementary integrating network of Fig. 2 was used for the present investigation.

The performance of the loop for various types of input signal will now be considered. In Section 3.1 the results will be derived from first principles together with a diagrammatic representation of loop operation. This will be followed by a direct theoretical derivation in Section 3.2.

(3) LOOP SUPPLIED WITH VARIOUS TYPES OF INPUT SIGNAL

(3.1) Qualitative Treatment

(3.1.1) Considerations Leading to the Static Loop Equation.

In the absence of an input signal, v_1 will be zero and hence v_2 and v_3 will also be zero. The oscillator O will operate at a frequency determined by the standing voltage V_3, say, at the input to the oscillator control C. Let the controlled oscillator frequency in general be $(f_i + f_g)$ and the 'free-running' frequency corresponding to the standing voltage V_3 be $(f_i + f_{g0})$.

Let an input signal of frequency f_s be applied to the loop. The balanced mixer M produces an output of frequency $(f_i + f_{g0} - f_s)$. If $(f_{g0} - f_s)$ is less than the half-bandwidth of the narrow-band amplifier N, the mixer output passes through N to form the discriminator input signal v_1 at a frequency $(f_i + f_{g0} - f_s)$. The feedback process between the output of the controlled oscillator and the input to the discriminator involves only a frequency comparison in M. The controlled oscillator signal provides the switching voltage for M so that the mixer output has a level dependent only on that of the input signal and contains the feedback information in its frequency properties. Thus any voltage gain between the mixer and the discriminator will not enter the loop-gain equation, and v_1 may be taken as the input signal level. Any shaping or time delay occurring between M and D must be accounted for. The gain/frequency characteristic of N may be considered normalized to its value at the frequency f_i and the gain transferred to the left of M (Fig. 1). Also, for large values of $(f_g - f_s)$ the transfer function of the mixer may be involved, but in all uses of the loop in the present application this action of the mixer was ignored since an amplifier N having a comparatively narrow bandwidth was used. The amplifier gain/frequency characteristic is shown in Fig. 3(*a*), and to a good approximation v_1 could be taken as constant when $(f_g - f_s)$ fell within the bandwidth and zero when it was greater than half the bandwidth.

Consider now the effect of the presence of the signal of frequency f_s at the mixer input, while the oscillator O runs at a frequency $(f_i + f_{g0})$. Let $(f_{g0} - f_s)$ be initially slightly less than the half-bandwidth of the amplifier N, as represented in Fig. 4. The signal v_1 will then produce an output v_2 such that

$$v_2 = c_1 v_1 (f_{g0} - f_s) \quad . \quad . \quad . \quad . \quad . \quad . \quad (1)$$

or

$$v_2 = k_1 (f_{g0} - f_s), \text{ say} \quad . \quad . \quad . \quad . \quad . \quad . \quad (2)$$

where

$$k_1 = c_1 v_1 \quad . \quad . \quad . \quad . \quad . \quad . \quad . \quad . \quad (3)$$

and c_1 (in volts per c/s) is a constant for the discriminator relating the voltage change at the output to unit frequency difference between the input frequency and the discriminator tuned frequency f_i. In the following discussion v_2 is taken as positive when f_g is greater than f_s. The results considered below indicate that for the values of $(f_g - f_s)$ encountered in practice and for the type of discriminator employed, k_1 may be assumed to be independent of $(f_g - f_s)$.

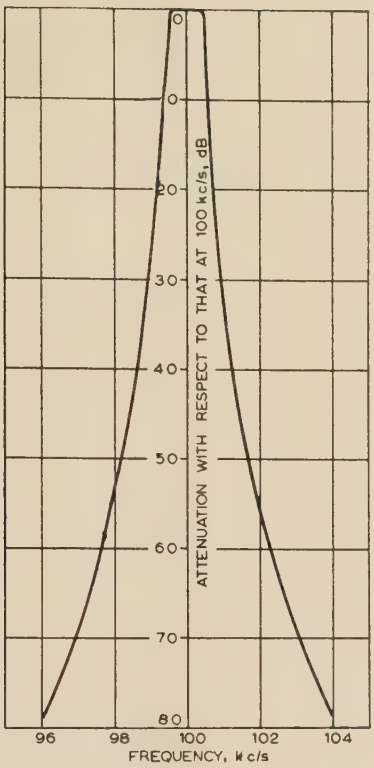
The voltage v_2 is applied to the integrating network S. The output from S, i.e. v_3, will initially fall at a rate v_2/C_1R_1, since the d.c. gain, $-A$, is made very large for reasons mentioned in Section 3.3.1. The oscillator control is arranged in such a way that a fall in input voltage causes the oscillator frequency to fall.

i.e.

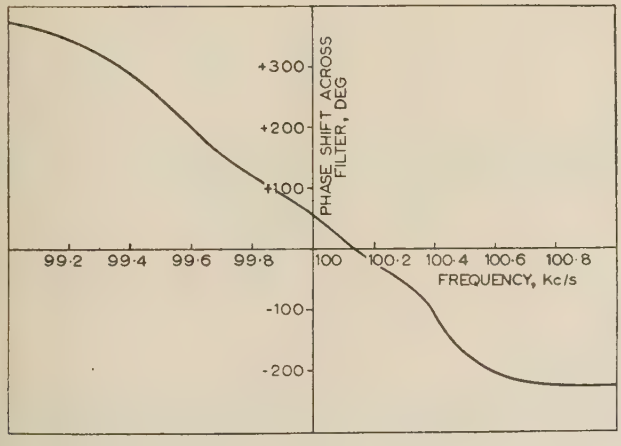
$$(f_{g0} - f_g) = -k_2 v_3 \quad . \quad . \quad . \quad . \quad . \quad . \quad (4)$$

where k_2 (in c/s per volt) is a constant for the oscillator and its control, relating the change in oscillator frequency to unit change in voltage v_3. Thus the presence of a fixed input signal frequency causes the loop to function in such a way that $(f_{g0} - f_s)$ is reduced. The frequency of the controlled oscillator will be constrained away from its free-running value until it acquires that frequency which, when mixed with the input signal frequency, produces just the required loop error voltage to constrain the oscillator frequency to this position. This process is known as 'pulling into lock'. When the loop is locked to a stationary input signal frequency, v_2 is constant, and

$$v_3 = -Av_2 \quad . \quad . \quad . \quad . \quad . \quad . \quad . \quad (5)$$



(*a*)



(*b*)

Fig. 3.—Characteristics of the narrow-band amplifier.
(*a*) Gain/frequency characteristic.
(*b*) Phase/frequency characteristic.

Hence $$(f_g - f_s) = \frac{f_{g0} - f_g}{Ak_1k_2} \quad . \quad . \quad . \quad . \quad . \quad (6)$$

combining eqns. (2), (4) and (5).

or $$\Delta f = \frac{f_{g0} - f_g}{Ak_3} \quad . \quad . \quad . \quad . \quad . \quad . \quad (7)$$

and $$\Delta f = \frac{f_{g0} - f_s}{1 + Ak_3} \quad . \quad . \quad . \quad . \quad . \quad . \quad (8)$$

where $$\Delta f = f_g - f_s \quad . \quad . \quad . \quad . \quad . \quad . \quad (9)$$

and $$k_3 = k_1k_2 \quad . \quad . \quad . \quad . \quad . \quad . \quad . \quad (10)$$

Δf will be called the 'frequency error' since the discriminator produces an error voltage proportional to $(f_g - f_s)$,

namely $$v_2 = k_1(f_g - f_s) = k_1\Delta f \quad . \quad . \quad . \quad . \quad (11)$$

Eqn. (2) was a particular case of eqn. (11) corresponding to the 'free-running' condition of the oscillator. Eqn. (8) is the static loop equation applicable to the loop locked to a stationary input signal frequency as represented in Fig. 4. The factor k_3A is the d.c. loop gain and is dimensionless.*

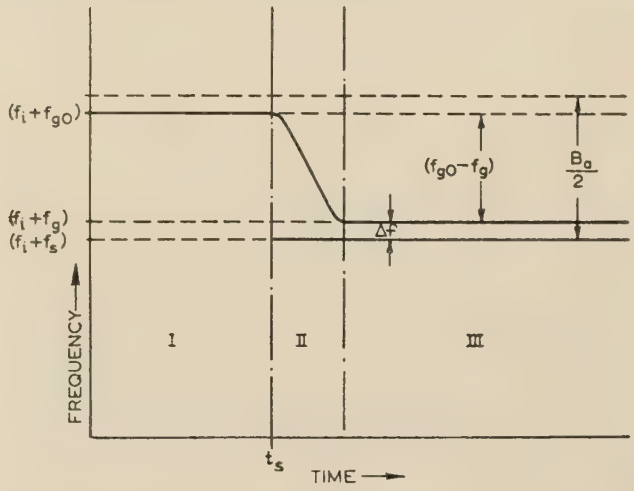


Fig. 4.—Representation of loop conditions before and after lock to a stationary input signal frequency.
Zone I: Controlled oscillator 'free-runs' at frequency $(f_i + f_{g0})$.
Zone II: Transient period during pull-in to lock.
Zone III: Stable, locked conditions.

In the present study a dynamic application of the loop was considered, since it was required to lock to and follow signals of changing frequency. The frequency errors for such cases will now be considered.

(3.1.2) Loop Supplied with a Step in Input Signal Frequency.

Commencing with the loop locked to a stationary input frequency, a step in this frequency may be applied. If the step is not of such magnitude that the frequency of the mixer output signal falls outside the bandwidth of the amplifier N (Fig. 1), the loop will pull into lock to the new frequency and the additional frequency error will be

$$\Delta f_1 = \frac{\pm f_s}{1 + Ak_3} \quad . \quad . \quad . \quad . \quad . \quad . \quad (12)$$

from eqn. (8), where $\pm f_s$ is the frequency step.

(3.1.3) Loop Supplied with a Step of Rate of Change of Frequency in the Input Signal.

In order that the loop may remain locked to an input signal of changing frequency, the frequency of the loop oscillator must change at such a rate that the frequency error does not exceed a certain value called the unlock threshold discussed in Section 4. Further, the frequency error at any instant will be of such a value that the corresponding voltage v_2 at the discriminator output, when acted upon by the shaping network, provides the position and rate of change of controlled oscillator frequency necessary to maintain locked conditions. This is achieved through the action of the shaping network as follows.

The voltage v_2 consists of two parts, one of which is directly amplified in the shaping network to provide a voltage $-Av_2$

* The d.c. loop gain for a phase-lock loop had the dimension of the reciprocal of time (see Reference 12).

which positions the frequency of the controlled oscillator according to eqns. (7) and (11). Considering Fig. 5, the directly amplified part Δf_{2a} is given by BC at time t, and from eqn. (8),

$$\Delta f_{2a} = \frac{f_{g0} - [F_s(t - t_0) + f_{g0}]}{1 + Ak_3}$$

$$= \frac{-F_s(t - t_0)}{1 + Ak_3} \quad . \quad . \quad . \quad . \quad . \quad . \quad (13)$$

Δf_{2b}, the second part of the frequency error, is represented by CD at time t in Fig. 5, and it produces a voltage $k_1\Delta f_{2b}$ at the

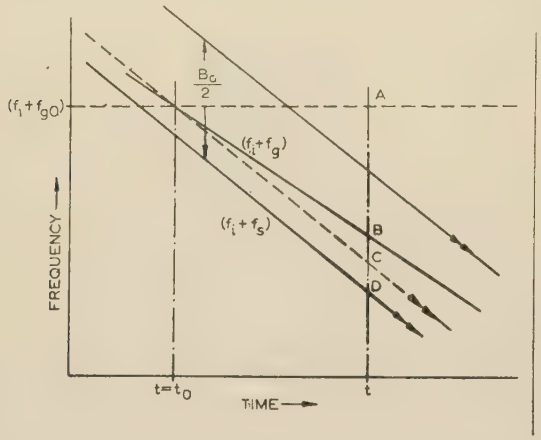


Fig. 5.—Representation of conditions when the loop tracks a signal whose frequency changes at a constant rate.

discriminator output according to eqn. (11). This causes the shaping network output voltage to fall at a rate $k_1\Delta f_{2b}/C_1R_1$, taking the network to act as a pure integrator since A is made large. The resulting rate of change of controlled oscillator frequency due to Δf_{2b} is identical to F_s, and hence

$$\Delta f_{2b} = -F_s\frac{C_1R_1}{k_3} \quad . \quad . \quad . \quad . \quad . \quad . \quad (14)$$

The sign of the errors is important. For the case shown, F_s is negative, and to produce this, CD must be positive so that f_g is greater than f_s. Further, since f_g is less than f_{g0} at the time t shown, BC must assume an additional positive value. In practice, A is made very large so that BC is much smaller than CD.

The total error Δf_2 which results when the loop tracks a signal of changing frequency is therefore

$$\Delta f_2 = \Delta f_{2a} + D\Delta f_{2b}$$

$$= \frac{-F_s(t - t_0)}{1 + Ak_3} + \frac{-F_sC_1R_1}{k_3}$$

In practice $Ak_3 \gg 1$, and

$$\Delta f_2 = \frac{-F_s}{k_3}\left[\frac{1}{A}(t - t_0) + C_1R_1\right] \quad . \quad . \quad . \quad (15)$$

Eqn. (15) and Fig. 5 show that the controlled oscillator frequency changes at a rate different from the rate of change of input signal frequency by an amount required to provide the position error. The position error changes with time but has fixed limits for any given working spectrum of input signal frequency.

(3.2) Theoretical Treatment

(3.2.1) The Loop Equation.

The derivation of an equation in terms of loop parameters, frequency error and input signal enables the performance of the loop to be determined for various types of input signal.

For practical values of the shaping-network time-constant used in the present work, the effect of the delay time in the narrow-band amplifier could be ignored by comparison. Taking the shaping network as the only reactive element in the loop, and considering Fig. 1, from eqn. (11),

$$v_2 = k_1\Delta f$$

$$v_3 = Y_s(p)v_2 \quad . \quad . \quad . \quad . \quad . \quad . \quad (16)$$

where $Y_s(p)$ is the transfer function of the shaping network given by

$$Y_s(p) = \frac{-1}{\frac{1}{A} + pC_1R_1} \quad . \quad . \quad . \quad . \quad (17)$$

when A is large.

Also, $(f_{g0} - f_g) = -k_2v_3$, from eqn. (4).

Combining eqns. (11), (16), (17) and (4),

$$\Delta f = \frac{1}{k_3}\left(\frac{1}{A} + pC_1R_1\right)(f_{g0} - f_g) \quad . \quad . \quad . \quad (18)$$

or, using eqn. (9),

$$\Delta f = \frac{\frac{1}{A} + pC_1R_1}{\frac{1}{A} + k_3 + pC_1R_1}(f_{g0} - f_s) \quad . \quad . \quad . \quad (19)$$

From eqn. (19), the general equation for the steady-state errors, following an input signal of type $f_s(p)$, will be given by

$$\underset{t\to\infty}{\Delta f} = \underset{p\to 0}{Lt}\left[\frac{-pf_s(p)\left(\frac{1}{A} + pC_1R_1\right)}{\frac{1}{A} + k_3 + pC_1R_1}\right] + \frac{f_{g0}}{1 + Ak_3} \quad . \quad (20)$$

(3.2.2) The Steady-State Error which Results when the Input Signal Frequency Changes by a Given Amount.

In this case, for an input-frequency step of magnitude f_s,

$$f_s(p) = \frac{f_s}{p}$$

and from eqn. (20), $$\Delta f_1 = \frac{f_{g0} - f_s}{1 + Ak_3} \quad . \quad . \quad . \quad . \quad . \quad (21)$$

This relation was derived from first principles in Section 3.1 as eqn. (8).

(3.2.3) The Error which Results when the Input Frequency Changes at a Constant Rate.

From eqn. (20), using $f_s(p) = F_s/p^2$,

$$\Delta f_2 = \underset{p\to 0}{Lt}\left[\frac{-F_s}{p}\left(\frac{\frac{1}{A} + pC_1R_1}{\frac{1}{A} + k_3 + pC_1R_1}\right)\right] + \frac{f_{g0}}{1 + Ak_3}$$

$$= \frac{-[F_s(t - t_0) + f_{g0}]}{1 + Ak_3} - F_s\frac{C_1R_1}{\frac{1}{A} + k_3} + \frac{f_{g0}}{1 + Ak_3}$$

Hence $$\Delta f_2 = \frac{-F_s(t - t_0)}{1 + Ak_3} - F_s\frac{C_1R_1}{\frac{1}{A} + k_3}$$

In practice $Ak_3 \gg 1$, and

$$\Delta f_2 = \frac{-F_s}{k_3}\left[\frac{1}{A}(t - t_0) + C_1R_1\right] \quad . \quad . \quad . \quad (22)$$

Eqn. (22) may be compared with eqn. (15) derived previously.

(3.3) Minimization of the Frequency Error

(3.3.1) Minimization of the Frequency Error in the Absence of Noise.

When a frequency-lock loop is to be used to track changes in the input signal frequency, it is necessary to design the loop so that the resulting frequency error is small for the type of signal-frequency changes involved. Some general statements may be made regarding the cases considered in the preceding discussion.

The error which results when the loop is locked to a signal of fixed frequency is given by eqn. (21). This error may be reduced by two procedures. First, the term Ak_3 in the denominator of eqn. (21) may be made large. Ak_3 represents the d.c. loop gain, namely

$$Ak_3 = Ac_1v_1k_2, \text{ from eqns. (3) and (10).}$$

The d.c. loop gain is usually most conveniently arranged to be large by increasing either A or v_1 or both.

Secondly, the free-running frequency of the oscillator may be adjusted so that f_s has the value f_{g0} at the centre of the spectrum of f_s over which the loop is required to track. This is accomplished by presetting V_3 accordingly, thereby ensuring that the maximum position error, which results when the oscillator frequency is constrained farthest from its free-running value, has been reduced to a minimum.

Eqn. (21) then approximates to

$$\Delta f_1 = \frac{f_{s0} - f_s}{Ak_3} \quad . \quad . \quad . \quad . \quad . \quad . \quad (23)$$

The conditions for minimization of the frequency error which results when the loop tracks a signal whose frequency changes at a certain rate may be drawn from eqn. (22). The time-varying part of this error is merely the integral of the error Δf_1, and this part may be made very small by the procedures outlined immediately above. The error Δf_2 then reduces to

$$\Delta f_2 = -F_s\frac{C_1R_1}{k_3} \quad . \quad . \quad . \quad . \quad . \quad (24)$$

$$= -F_sT, \text{ say} \quad . \quad . \quad . \quad . \quad (25)$$

where

$$T = \frac{C_1R_1}{k_3} \quad . \quad . \quad . \quad . \quad . \quad . \quad (26)$$

This error corresponds to the 'velocity lag' of a position-control servo mechanism using servo-mechanism control terminology, and the factor $1/T$ is the a.c. loop gain.

Δf_2 may be made as small as desired by increasing k_3 and reducing the shaping-network time-constant.

(3.3.2) Minimization of the Frequency Error when the Input Signal is in the Presence of Noise.

The input signals used in the present investigation were in the presence of a high level of random noise of wide bandwidth. This noise passed through the loop mixer and narrow-band amplifier and produced a mean frequency error at the output of the loop discriminator. The total error at the discriminator output comprises the frequency error formed when tracking signal frequency changes and the mean noise error. The mean-square noise error may be shown to be proportional to a loop noise bandwidth B_n, where[5,9,12]

$$B_n \propto \frac{1}{T} = \frac{k_3}{C_1R_1} \quad . \quad . \quad . \quad . \quad . \quad (27)$$

It is apparent that the reduction of the error Δf_2 by increasing the a.c. loop gain is incompatible with the reduction of the noise error. For a given signal/noise ratio and tracking rate, an optimum loop gain will exist for which the total error is made a minimum. When the signal/noise ratio is low, a compromise must be struck between making the loop of sufficiently wide bandwidth to track signal-frequency changes and making it insensitive to the noise. This is the basis of the theory given in Section 5, where the nature of the loop tracking performance in noise is examined, but first a criterion of this performance must be introduced.

(4) THE MAXIMUM RATE OF FOLLOW: A CRITERION OF TRACKING PERFORMANCE

In the absence of noise, the steady-state frequency error may be increased up to a limit where the loop becomes unlocked, and this limit is well defined. When the loop is tracking a changing signal frequency, unlock will occur if the steady-state frequency error exceeds the limit, to be termed here the 'unlock threshold'.

When noise is present, the total frequency error is composed of a discrete error due to tracking the signal-frequency changes and a randomly-varying noise error. The largest signal error which may exist before the total frequency error exceeds the unlock threshold is then dependent upon the statistical properties of the noise. The maximum signal error was therefore determined by a statistical method as described in Section 7.2, where a factor called the 'mean time-to-unlock' is introduced. Since this error prescribes the rate of change of controlled oscillator frequency, according to eqn. (25) it represents the maximum rate of change of signal frequency which the loop is capable of tracking for a particular level of noise and particular loop constants. This maximum rate was called the maximum rate of follow, $\hat{F}_s$.

(5) THEORY OF THE LOOP TRACKING PERFORMANCE WHEN THE INPUT SIGNAL IS IN THE PRESENCE OF NOISE

The variation of the maximum rate of follow with changes in signal/noise ratio and loop parameters may be derived as follows from a consideration of the unlock threshold and the signal and noise errors.

Let Δf_t be the total frequency error representing the unlock threshold for a mean time-to-unlock greater than 40 sec. A certain proportion of this total error will be due to the noise. Plotting T, the inverse of the a.c. loop gain, as the abscissa, let the noise error Δf_n vary inversely with T as shown in Fig. 6, curve (*a*). The maximum allowable steady-state signal error $\Delta\hat{f}_2$ at a particular value of T is then given by $(\Delta f_t - \Delta f_n)$. An important feature of the rectangular hyperbola for Δf_n in Fig. 6 is that it crosses the unlock threshold at a critical value of T, T_c say. There is therefore a finite value of T for which the noise error alone exceeds the unlock threshold, so that zero signal error can exist for values of T less than T_c.

If $\Delta\hat{f}_{2r}$ is the maximum signal error when $T = rT_c$, as indicated in Fig. 6,

$$\Delta f_nT = (\Delta\hat{f}_2 + \Delta f_n - \Delta\hat{f}_{2r})rT_c$$

and

$$\Delta f_nT = (\Delta\hat{f}_2 + \Delta f_n)T_c$$

Hence

$$\Delta\hat{f}_2 = \frac{r}{r-1}\left(1 - \frac{T_c}{T}\right)\Delta\hat{f}_{2r} \quad . \quad . \quad . \quad . \quad (28)$$

The value of Δf_2 which results when the loop tracks a changing signal frequency is given by eqn. (24) or (25). Curve (*b*) is

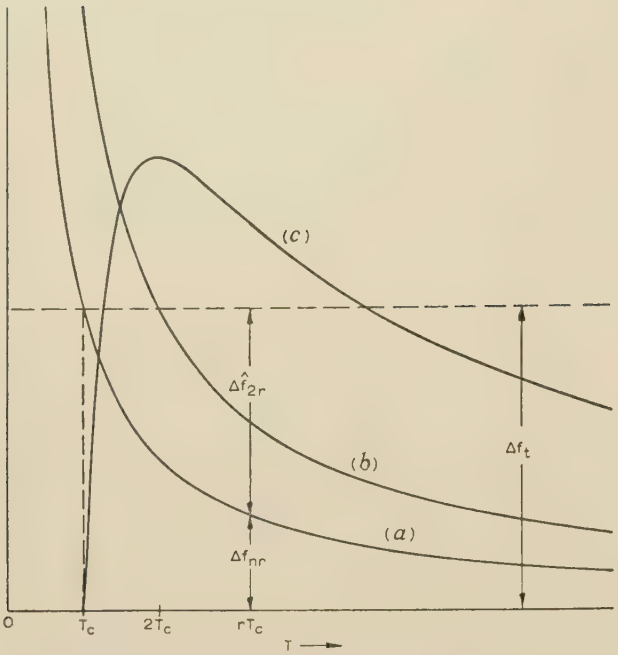


Fig. 6.—Derivation of the maximum rate-of-follow curve.

plotted in Fig. 6 showing the variation with T of the rate of follow per unit frequency error according to eqn. (25), namely

$$\frac{F_s}{\Delta f_2} = \frac{-1}{T} \quad . \quad . \quad . \quad . \quad . \quad . \quad . \quad (29)$$

Curve (*b*) is also a rectangular hyperbola.

At a particular value of T, the maximum rate of follow $\hat{F}_s$ is given by the product of the maximum allowable error $\Delta\hat{f}_2$ and the rate of follow per unit error. At each value of T the ordinate of curve (*b*) is multiplied by $\Delta\hat{f}_2$ to give curve (*c*), which shows the variation of $\hat{F}_s$ with T. The equation for curve (*c*), obtained from eqns. (28) and (29), is

$$\hat{F}_s = \frac{-r}{r-1}\frac{1}{T}\left(1 - \frac{T_c}{T}\right)\Delta\hat{f}_{2r} \quad . \quad . \quad . \quad . \quad (30)$$

$\hat{F}_s$ has a maximum (negative) value given by $d\hat{F}_s/dT = 0$ with $d^2\hat{F}_s/dT^2$ positive. It follows from eqn. (30) that $\hat{F}_s$ has a maximum value when $T = 2T_c$. With $T = 2T_c$ and hence $r = 2$, eqn. (28) then shows that $\Delta\hat{f}_2$ forms half the total error as determined by the unlock threshold. For a particular signal/noise ratio, the condition $T = 2T_c$ provides optimum tracking performance. Curve (*c*) illustrates the compromise required between the large values of noise error for small values of T and the small values of rate of follow per unit error for high values of T.

If $\hat{F}_{sv}$ is the maximum rate of follow when $T = 2T_c$, $r = 2$ and, from eqn. (29),

$$(\Delta f_{2r})_{r=2} = -\hat{F}_{sv}2T_c \quad . \quad . \quad . \quad . \quad . \quad (31)$$

Eqn. (30) may then be expressed as

$$\hat{F}_s = 4\frac{T_c}{T}\left(1 - \frac{T_c}{T}\right)\hat{F}_{sv} \quad . \quad . \quad . \quad . \quad (32)$$

Eqn. (32) provides the relation between $\hat{F}_s$ and T for a fixed signal/noise ratio.

It is now of interest to determine how the maximum follow rate is affected by alterations in the signal/noise ratio. When this changes, T_c will vary in such a way that for high signal/noise ratios a lower value of T (or higher noise bandwidth B_n) will be reached before Δf_n exceeds the unlock threshold. The variation of T_c with signal/noise power ratio is taken to be hyperbolic, i.e. if T_c' and S' are particular values of T_c and the signal/noise ratio S,

$$T_c = \frac{S'}{S}T_c' \quad . \quad . \quad . \quad . \quad . \quad . \quad . \quad (33)$$

From eqns. (29), (31) and (33) it follows that, if $\hat{F}_{sv}'$ is the value of $\hat{F}_{sv}$ corresponding to T_c' and S',

$$\hat{F}_{sv} = \frac{S}{S'}\hat{F}_{sv}' \quad . \quad . \quad . \quad . \quad . \quad . \quad . \quad (34)$$

Using eqns. (33) and (34), eqn. (32) becomes

$$\hat{F}_s = 4\frac{T_c'}{T}\left(1 - \frac{S'}{S}\frac{T_c'}{T}\right)\hat{F}_{sv}' \quad . \quad . \quad . \quad . \quad (35)$$

This equation provides the complete ($\hat{F}_s$, T) plane, and a series of curves is plotted in Fig. 7 showing the variation of $\hat{F}_s$ with T for each of a range of values of signal/noise ratio.

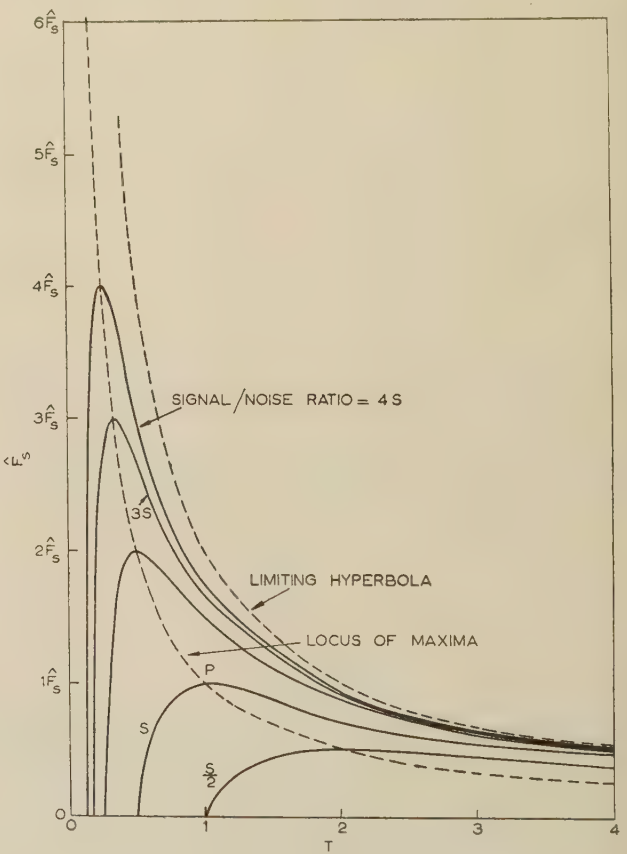


Fig. 7.—The complete ($\hat{F}_s$, T) plane.

The variation of S and T for each of a range of values of $\hat{F}_s$ is also of concern. Considering eqn. (35), with $\hat{F}_s$ constant, and if S_v is the signal/noise ratio when $T = 2T_c$,

$$1 = 4\frac{T_c}{T}\left(1 - \frac{S_v}{S}\frac{T_c}{T}\right)$$

or

$$S = \frac{T_c}{T\left(1 - \frac{T}{4T_c}\right)}S_v \quad . \quad . \quad . \quad . \quad . \quad (36)$$

The complementary set of curves for fixed values of $\hat{F}_s$ and variable signal/noise ratio derived from eqn. (36) is shown in

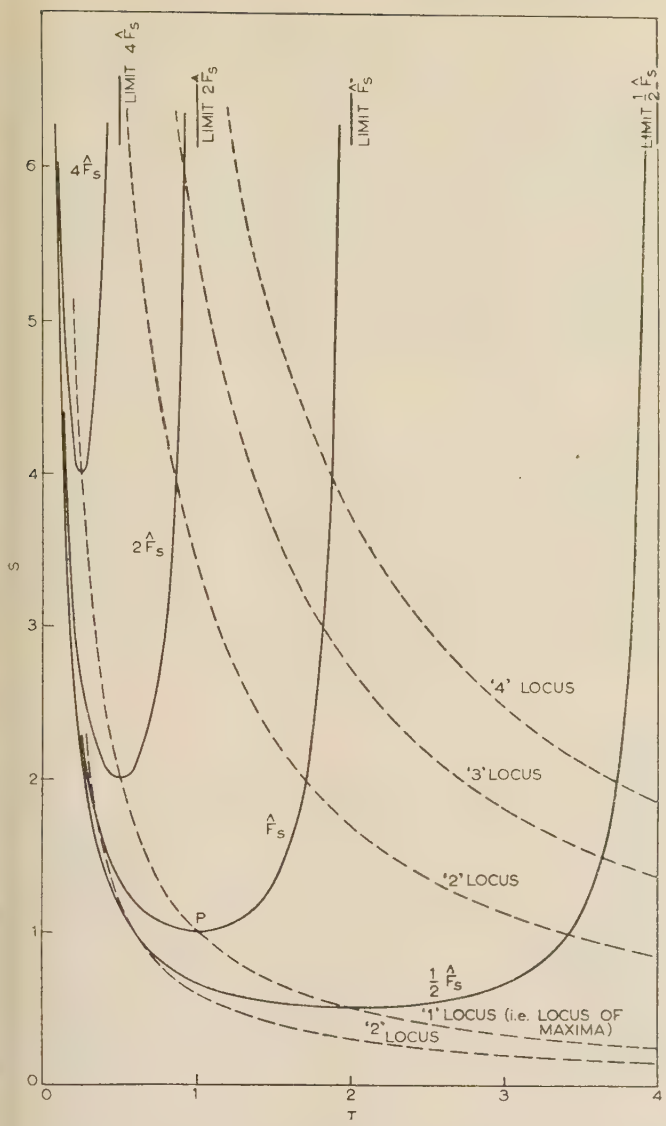


Fig. 8.—The complete (S, T) plane.

The '1', '2', '3' and '4' loci are rectangular hyperbolae and each curve cuts these loci at values of S 1, 2, 3 and 4 times that at its vertex.

Fig. 8. The lowest signal/noise ratio for which the loop will follow at a given rate of change of frequency corresponds to the tangential point P in Fig. 7 or the minimum value of P in Fig. 8.

A plot corresponding to that of Fig. 7, but with $1/T$ or a.c. loop gain as abscissa, is shown in Fig. 9. The maximum rate-of-follow curves appear as parabolae with the locus of their vertices on a straight line passing through the origin.

(5.1) Zone of Operation in the ($\hat{F}_s$, T) Plane

It is of interest to determine whether the curves representing $\hat{F}_s$ for all values of S with T fill the complete zone between the $\hat{F}_s$ and T axes. As may be seen most easily from the series of parabolae plotted in Fig. 9, eqn. (35) may be rewritten in the following manner:

Let
$$\frac{1}{T} = x$$
$$\hat{F}_s = y$$
$$\hat{F}_{sv} = b$$
and
$$\tan\alpha = 2\hat{F}'_{sv}T'_c$$

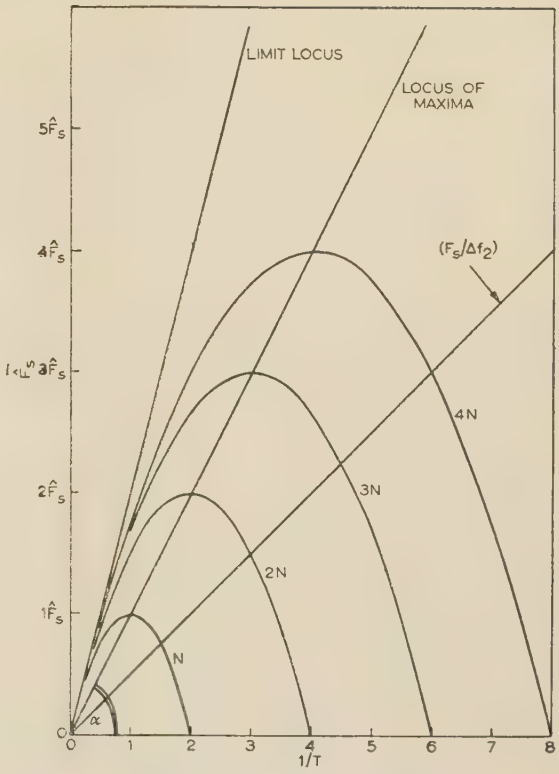


Fig. 9.—The ($\hat{F}_s$, $1/T$) plane.

α is the angle between the locus of maxima and the $1/T$ axis as shown in Fig. 9.

From eqn. (34), eqn. (35) becomes

$$x^2 - 2bx\cot\alpha + by\cot^2\alpha = 0 \quad . \quad . \quad . \quad (37)$$

The corresponding form of eqn. (35) is

$$bx^2 - 2abx + a^2y = 0 \quad . \quad . \quad . \quad . \quad . \quad (38)$$

where
$$\frac{1}{2T_c} = a$$

Differentiating eqn. (37) with respect to x,

$$2x - 2b\cot\alpha + b\frac{dy}{dx}\cot^2\alpha = 0$$

When $x = 0$, $\dfrac{dy}{dx} = 2\tan\alpha$.

Hence there is a limit line of slope twice that of the locus of maxima on Fig. 9, and the zone of loop operation is confined to the area between the limit line and the $1/T$ axis. The corresponding limit locus shown in Fig. 7 is given by a hyperbola of constant twice that for the locus of maxima. In particular, the limit locus represents the variation of $\hat{F}_s$ with T when noise is absent.

(6) DESCRIPTION OF THE LOOP USED

The various component parts of the frequency-lock loop used in the present series of measurements had the following general properties.

The balanced mixer M, of Fig. 1, was tuned to 100kc/s and was switched by the signal from the controlled oscillator, which was a multivibrator. The narrow-band amplifier N following the mixer had the gain and phase characteristics shown in Fig. 3, and a centre frequency of 100kc/s.

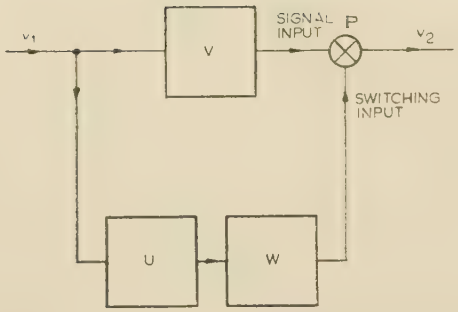


Fig. 10.—The units comprising the frequency discriminator.
U : Single-stage tuned circuit.
W : Wide-band amplifier.
V : Variable phase-shifter.
P : Phase-sensitive detector.

The discriminator consisted of the units shown in Fig. 10. The signal v_1 from N was passed jointly to a variable phase-shifter V and a single-stage tuned circuit U. The tuned circuit was of variable bandwidth B_c, and had a centre frequency of 100 kc/s. The output from U was amplified in a wide-band amplifier W to provide the switching signal of a phase-sensitive detector P. The signal from the variable phase-shifter formed the input signal for this detector whose output v_2 was passed to the shaping network. The phase-shifter V was preset so that v_2 was zero when v_1 had a frequency of 100 kc/s. When the frequency of v_1 differed from 100 kc/s, the phase of the switching signal for P changed according to the phase-shift/frequency characteristic of U. The change was detected by P, which provided a d.c. output k_1 (in volt per c/s). For a particular signal level v_1 and bandwidth B_c, k_1 could be assumed constant over the range of frequency errors encountered in the measurements.

The shaping network is shown in Fig. 2, together with typical values of C_1, R_1 and A. The circuit was a Miller integrator with cathode-following feedback. R_1 included the output resistance of P.

The output v_3 was transmitted by the oscillator control to the grid-leaks of the multivibrator. By this means v_3 controlled the aiming potential of the multivibrator and hence its frequency. The factor k_2 (in kc/s per volt) was virtually constant in the working frequency range of the oscillator.

(7) SETTING-UP PROCEDURE AND MEASURING TECHNIQUE

(7.1) Setting-up Procedure

With the loop opened at the shaping-network input, the discriminator was first set up by tuning the discriminator circuit U to 100 kc/s and adjusting the phase-shifter V for zero output v_1. The loop was then closed and the loop locked to a stationary input signal frequency in the absence of noise. The frequency of the signal to the discriminator was measured, and the difference between this and 100 kc/s was required to be small. A measurement of this difference at both ends of the range of signal frequencies used provided a check to ensure that the d.c. loop gain was sufficient to make the position error negligibly small. The loop was then in a satisfactory condition for use with noise-free signals, but a further procedure was needed in the presence of noise, as follows.

The input signal was removed and wide-band noise applied in its place. Owing to the inevitable slight asymmetry in the amplitude/frequency characteristic of the amplifier N, its mean frequency differed slightly from 100 kc/s and a small residual error was observed at the discriminator output. This was reduced to zero by altering the centre frequency of the tuned circuit U to this mean frequency.

(7.2) A Technique for Measurement of the Maximum Rate of Follow

The maximum rate of follow is determined by the signal error at unlock, as described in Section 4, and the following method uses this property.

First, the noise is considered to be absent at the input. The junction of R_1 and C_1 in Fig. 2 was connected via a resistor R_L to a variable direct-voltage source external to the loop. The junction of R_1 and C_1 was very near the virtual earth potential since the d.c. gain A was large and $-j/\omega C_1(A+1)$ was small. Independent currents could be fed into the virtual earth from the discriminator output voltage v_2 and from the direct-voltage source v_L to charge C_1. The current charging C_1 was zero when

$$\frac{v_L}{v_2} = \frac{R_L}{R_1} \quad . \quad . \quad . \quad . \quad . \quad . \quad . \quad (39)$$

to a good approximation, where R_1 included the output resistance of the discriminator. With the loop locked to a fixed input signal frequency and $v_L = 0$, v_2 consisted of a very small position error according to eqn. (21). On applying a small voltage v_L, v_2 became transient and settled to a value determined by eqn. (39). v_2 and v_L then effectively charged C_1 at equal and opposite rates, so keeping the controlled oscillator frequency fixed, but the frequency error was increased by an amount

$$\Delta f = \frac{v_L R_1}{k_1 R_L} \quad . \quad . \quad . \quad . \quad . \quad . \quad . \quad (40)$$

from eqns. (11) and (39).

Δf was measured by recording the frequency of the input signal to the discriminator before and after application of the voltage v_L. The input signal was then removed and v_2 became zero, but v_L remained and the resulting rate of change of controlled oscillator frequency F_g was measured. If the loop had been tracking an input signal frequency changing at a rate F_s, an error Δf as given by eqn. (40) would have been produced when F_s had the value F_g.

This procedure was extended to a measurement of the maximum rate of follow $\hat{F}_s$, by measuring the value of F_g corresponding to the value of v_L which just caused the loop to become

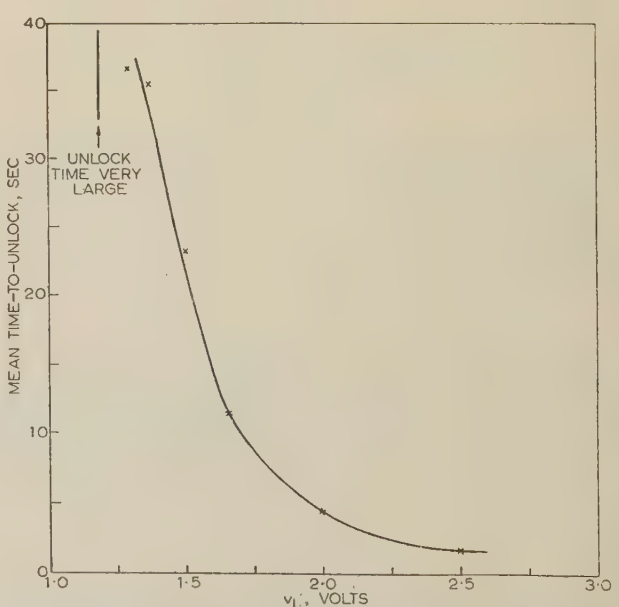


Fig. 11.—Variation of the mean time-to-unlock with external error voltage, v_L.

unlocked. The frequency error corresponding to this particular value of v_L was the unlock threshold.

When noise is present at the input, the above method may be applied, but first the signal/noise ratio must be defined. The signal and noise powers were measured separately at the input to the loop discriminator. The signal power P_s was measured in the absence of noise with the loop locked and with $v_1 = 0$. The noise power P_N was measured with the signal level zero. When both signal and noise were present, the total power measured was not quite the sum of P_s and P_N owing to the jitter of the controlled oscillator frequency. This jitter reduced the mean signal power from the value P_s, since it was an appreciable fraction of the bandwidth of the narrow-band filter and the amplitude characteristic of the latter was not perfectly flat in the pass band [see Fig. 3(*a*)].

A particular signal/noise ratio having been set up an external voltage v_L was applied at a noted time, and the time which elapsed before unlock occurred was observed. The elapsed time was called the 'time-to-unlock', t_u. Owing to the random nature of the noise error it was necessary to measure t_u about a dozen times, for each setting of v_L, in order to obtain a mean value of t_u accurate to 10%. The results are shown in Fig. 11, which indicate that v_L is asymptotic to a value approaching that for a mean time-to-unlock of 40 sec. At large values of v_L, t_u was indistinct since it became of the same order as the time the loop took to complete transient conditions following the application of v_L. The curve of Fig. 11 may be regarded as showing the variation of the probability of the total error not exceeding the unlock threshold with signal error. (The effect could be likened to the case of the probability of electrons escaping from the 'potential well' of a metal by thermionic emission.)

The maximum rate of follow was taken as that value of F_g which was produced by a signal error giving a mean unlock time greater than 40 sec. The measurement was repeated for both senses of sweep and the mean recorded, and it is estimated that the results for $\hat{F}_s$ are in error by maxima of $\pm 5\%$.

The position error was negligibly small for the rate-of-follow measurement, since the d.c. loop gain was about 2300 and f_s was made equal to f_{g0}.

(8) RESULTS

(8.1) The Maximum Rate of Follow over a Range of Signal/Noise Ratio and A.C. Loop Gain

The maximum rate of follow was measured for a series of signal/noise ratios and over a range of a.c. loop gain. The combined results are given in Fig. 12, and each point on the curves shown was determined by the process of repetitive trials outlined in Section 7.2.

A series of curves is also plotted in Fig. 12 using the theory outlined above [eqn. (35)] and may be compared with the practical results obtained. Theoretical and practical results for the variation of T_c with signal/noise ratio [eqn. (33)] are shown in Fig. 13. The theoretical curves appear to give agreement with the results obtained, within the expected experimental error.

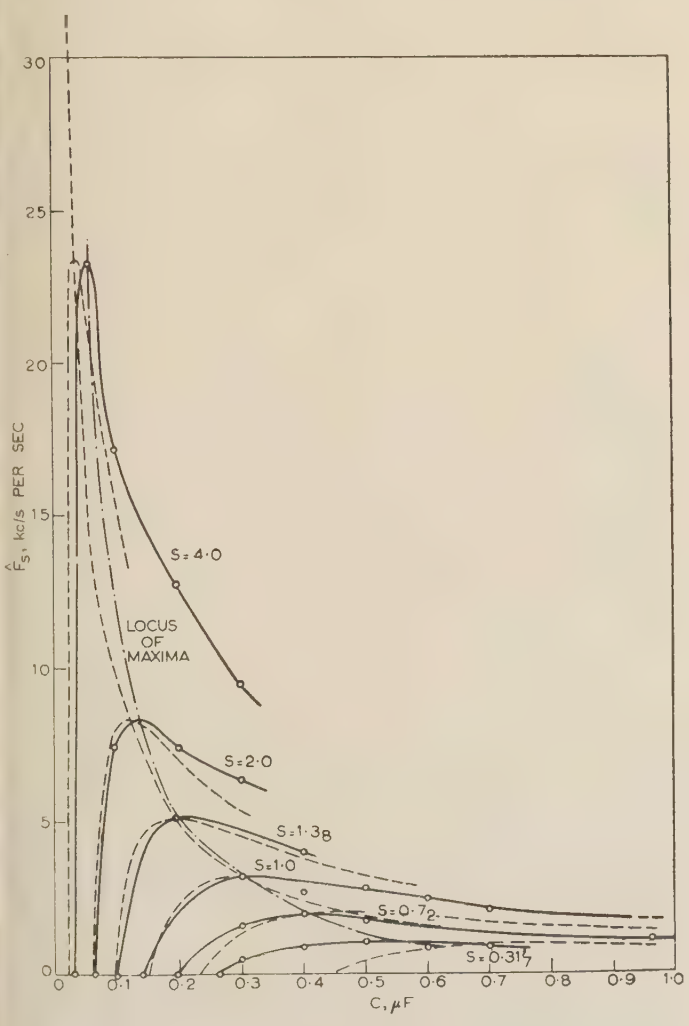


Fig. 12.—Combined theoretical and practical results.

$B_c = 462$ c/s, mean $t_u \ll 40$ sec.
$k_1 = 8{\cdot}13$ volt per kc/s, $k_2 = 2{\cdot}20$ kc/s per volt.
$R_1 = 1{\cdot}025$ megohms.
i.e. $T = 0{\cdot}0572\,C_1$ sec. C_1 is in microfarads.
S measured in 400 c/s bandwidth.

———— Practical.
— — — — Theoretical.

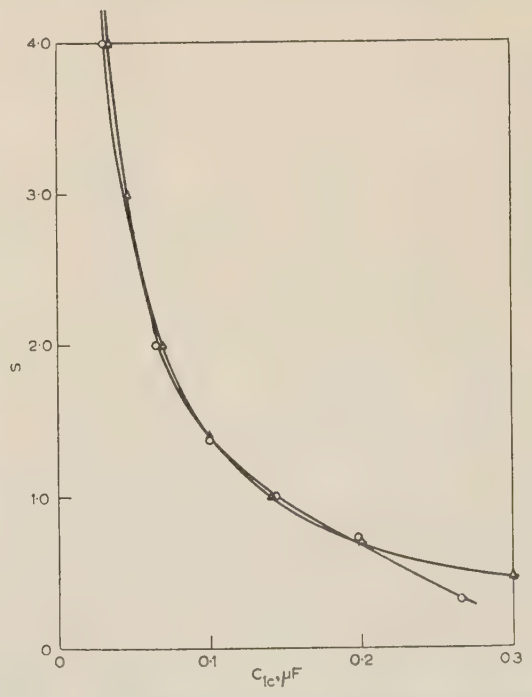


Fig. 13.—The variation of signal/noise ratio with critical capacitance C_{1c}.

Loop parameters are as shown in Fig. 12.
$T_c = 0{\cdot}0572\,C_{1c}$ sec. C_{1c} is in microfarads.

○ Practical.
△ Theoretical.

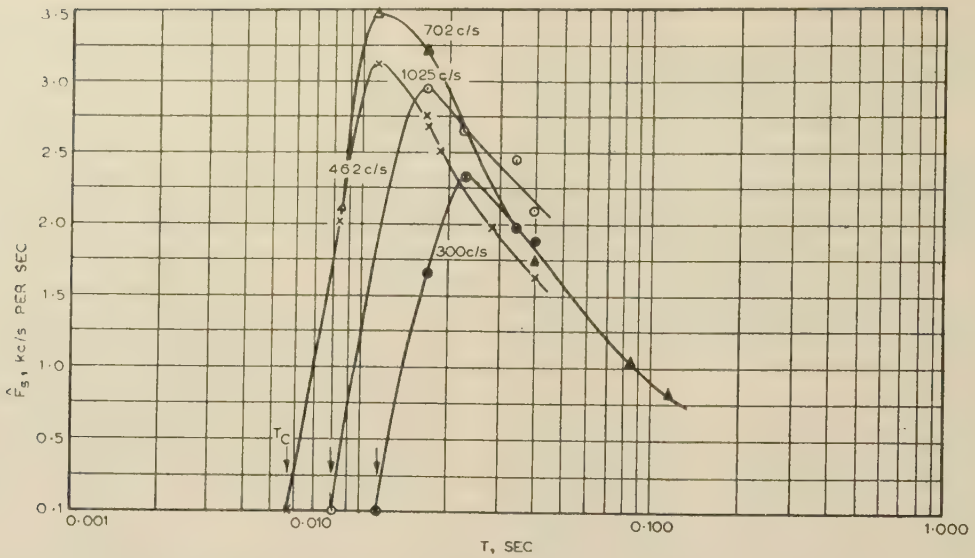


Fig. 14.—Maximum rates of follow over a range of discriminator bandwidths.

$S = 1{\cdot}00$ in 400 c/s bandwidth.
B_c:
× 1 025 c/s.
△ 702 c/s.
⊙ 462 c/s.
⊗ 300 c/s.

(8.2) Further Results

(8.2.1) The Effect of Varying the Discriminator Bandwidth.

Some additional measurements of the maximum rate of follow were made over a range of bandwidths of the discriminator filter, B_c, and are shown in Fig. 14 for a signal/noise ratio of unity in 400c/s bandwidth at the discriminator input. The effect of varying B_c on the maximum rate of follow at the optimum value of T is shown in Fig. 15. This curve has a shallow maximum at

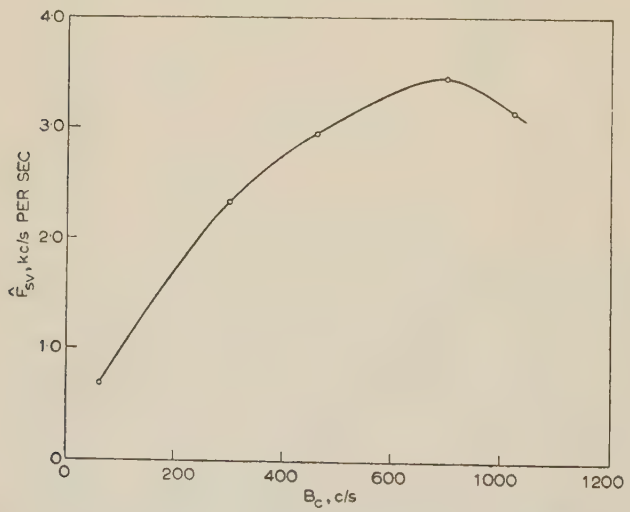


Fig. 15.—Maximum rate of follow at optimum T versus discriminator bandwidth.

a bandwidth of 750c/s and falls to low values for small bandwidths when the unlock threshold becomes small. The values of T shown in Figs. 14 and 15 include the effect of the change of k_1 with changing B_c.

(8.2.2) The Effect of Automatic Optimization of Loop Parameters with Changing Signal/Noise Ratio.

When the input signal/noise ratio varies widely and the a.c. loop gain is fixed, the loop will provide the optimum maximum rate of follow at only one particular signal/noise ratio. In Fig. 16, curve (*a*) shows the variation of the maximum rate of

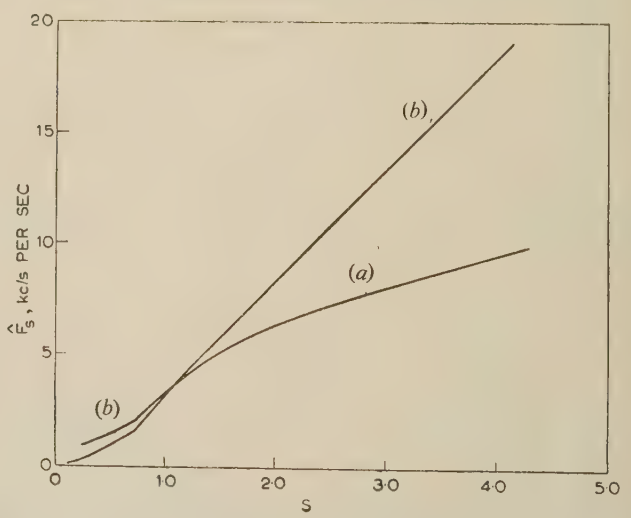


Fig. 16.—A typical comparison of loop performance for fixed and automatically varying parameters.

follow with the signal/noise ratio for the loop optimized at unity in 400c/s bandwidth. Curve (*b*) shows the variation that was obtained when T was varied with the signal/noise ratio by auxiliary means according to eqn. (33).

(8.3) The Use of more Complicated Networks

The results of the work carried out in the present investigation have involved a single integrator as the shaping network in the feedback loop. More complicated shaping networks might be used in some cases. A compromise would be required between the reduction of the noise bandwidth and the time taken to complete transient conditions at initial lock and after a change in the rate of follow.

(9) CONCLUSIONS

It has been shown that the tracking performance of a frequency-lock loop can be represented by a factor called the maximum rate of follow. This factor has been determined theoretically from an analysis of the loop operation in terms of loop parameters and the signal/noise ratio. A technique used for the measurement of the maximum rate of follow is described: when the input signal was in the presence of noise, a mean time-to-unlock greater than 40sec was required. The results provide general agreement with the theory within the expected experimental error of $\pm 5\%$.

(10) ACKNOWLEDGMENTS

The acknowledgments of the author are due to the Chief Engineer of the English Electric Company for permission to submit the paper for publication.

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DISCUSSION ON

'ACOUSTICS OF LARGE ORCHESTRAL STUDIOS AND CONCERT HALLS'*

Mr. K. L. Rao (*communicated*): Since 1939 I have had very few opportunities of listening to full-scale western orchestras of classical types; I have, however, listened to radio broadcasts of live as well as recorded pieces of this type of music. I am inclined to agree with the authors' concluding remarks in the last two paragraphs of Section 10. As they state, aural assessment and aesthetic judgment are the ultimate criteria for good or bad acoustics of this type of hall.

I have listened a great deal to the following music performances:

BBC concert series and many of their orchestral programmes from the Albert Hall and other places.

Netherlands Sunday concerts mostly emanating from the Amsterdam Concert Hall.

I have come to the conclusion that the best is from the Scottish Orchestra which comes from Glasgow. Next best is the Amsterdam hall. Third is the Albert Hall after it underwent renovations some time in 1949–50. The Royal Festival Hall programmes, somehow, did not come up to standard. After reading the paper the reason becomes clear.

I have also come to the conclusion that the old type of rectangular hall has given the best results. The modern design 'streamlines' everything, as it were. The reason may be that the material and the architecture give the necessary reverberation and at the same time the required diffusion from the broken surfaces of the architectural designs. I am reminded of the music performances held in temple halls in this country in the earlier days. The results were excellent indeed. (The music, of course, was Indian.) The rectangular temple hall, with its expansive area, high roof and carved pillars and walls appears to give the necessary 'life' and distribution of sound without having resort to artificial means. This type of hall appears to give the desired features of 'adequate scattering, absence of deliberate reinforcement of direct sound, ample height and an adequate reverberation time', to put it in the authors' words. In this connection I am also reminded of the organs played during church services, which are quite often live programmes. The organ comes out in all its glory. I had wondered why churches and temples were supposed to be acoustically bad, but in fact were not. After reading the paper many of the points become clear.

Messrs. T. Somerville and **C. L. S. Gilford** (*in reply*): As regards the concerts broadcast from Glasgow, it should be realized that the majority of these come from Glasgow Studio 1, mentioned on page 92, the orchestra being the B.B.C. Scottish Orchestra. There is another orchestra in Glasgow, the Scottish National Orchestra, which normally performs at public concerts in St. Andrew's Hall, but broadcasts are very infrequent and consequently it is unlikely that Mr. Rao has heard them. This orchestra also often broadcasts from Glasgow Studio 1 and it therefore appears likely that it is this studio which he is comparing with the Amsterdam Concert Hall and the Royal Festival Hall.

His general support for our views is gratifying.

* Somerville, T., and Gilford, C. L. S.: Paper No. 2297 R, March, 1957 (see 104 B, p. 85).

DISCUSSION ON

'THE CAPACITANCE BETWEEN DIODE ELECTRODES IN THE PRESENCE OF SPACE CHARGES'*

Mr. I. A. Harris (*communicated*): The statement in the paper that the equation $I = dD/dt$ for the total current in an ideal planar diode is incomplete because of the presence of the electrodes with their charges, is incompatible with eqns. (13)–(21). For, on applying eqn. (16) to the total current between parallel plane electrodes of sufficient extent to render edge effects negligible, $v = dx/dt$ and $\rho = \nabla D_s = \partial D_s/\partial x$, and on noting that $\nabla D_b = 0$ in the space (i.e. $\partial D_b/\partial x = 0$) we can write $\rho v = \partial D/\partial x \,.\, dx/dt$. Therefore

$$I = \frac{\partial D}{\partial x} \cdot \frac{dx}{dt} + \frac{\partial D}{\partial t}$$

where $D = D_s + D_b$, is valid throughout the space between the electrodes, provided the velocity is single valued at any plane in the space. For this equation to be limited to apply to D_s alone, $\partial D_b/\partial t$ would have to be identically zero, an absurd requirement when the electrode potential-difference has an alternating component.

It follows that the expression $I = k\bar{x}$ is valid, resulting in the expression $\frac{3}{5}\omega C$ for the susceptance to small signals of an ideal diode with a space-charge-limited current, where C is the capacitance in the absence of space charges. The value found for this susceptance in the paper is zero. Experimentally,† provided the space-charge-limited current density is high enough to move the potential minimum relatively close to the cathode, the value of the capacitance observed does not differ widely from $\frac{3}{5}C$.

I suggest that the discrepancy arises because Dr. Bull's calculation compares two slightly different steady states of the diode and takes no account of the inertia of the moving charges under the influence of an applied voltage that varies with time. The solution of the above equation, on the other hand, automatically takes this into account. In spite of some opinions expressed in the past, it is certain now that the change in susceptance (or equivalent capacitance) due to the presence of moving space charges is a time-of-flight effect, even at the lowest frequencies.

Dr. C. S. Bull (*in reply*): Mr. Harris takes $\nabla D_b = 0$ everywhere, so that for the planar case $\partial D_b/\partial x = 0$ everywhere, and so obtains Llewellyn's equation. On the other hand, I state if I is the total current between the electrodes we must make $\nabla \,.\, I = 0$ everywhere. Since I did not write the paper with Mr. Harris's objections in view, let us rewrite eqn. (13) as follows to define I:

$$I = \nabla \times H = J + \frac{\partial D}{\partial t}$$

The process of taking the divergence of this equation for the space between the electrodes is carried out in the paper, and will not be repeated here. The equation certainly satisfies my condition $\nabla \,.\, I = 0$ in the space between the electrodes.

At the surface of one of the electrodes $\nabla \,.\, D_b$ is no longer zero but becomes infinite for an infinitesimal distance. On going through the surface charge D_b becomes zero. The displacement current $\partial D_b/\partial t$ which exists between the electrodes can be replaced inside them by a conduction current of equal magnitude. This current will account for the accumulation of charge on the surface. The remaining portion of the current I which exists between the electrodes can be replaced inside them by a conduction current equal to $(\partial D_s/\partial t + \rho v)$. We have now produced a state of affairs described throughout the space between the electrodes and through the electrodes surfaces into the metal in which the total current is everywhere constant, i.e. $\nabla \,.\, I = 0$ everywhere.

(It should be noticed that the discontinuity in the value of D_b at the surface is not an essential mathematical feature of the situation. It can be removed by proceeding as above for many very small elements of current and making the point of discontinuity for each element occur at slightly different places over a very small region near the surface of the electrodes. This procedure will meet the physical needs of the case and at the same time preserve the condition that the divergence is everywhere zero.)

The main purpose of my paper was to show that the capacitance between two electrodes can be calculated if we assume that the current going into one is the same as that coming out of the other; in other words, by making the divergence of the total current zero. A simple case was worked out, and is now seen to be justified. On the other hand, Mr. Harris cannot consider by means of his equation the situation in the thin region on or near the surface of the electrodes at which the surface charges reside, since there $\nabla \,.\, D_b$ is not zero. Consequently he cannot draw any conclusions pertinent to the flow of current to actual electrodes. I contend, therefore, that the condition $\nabla \,.\, I \equiv 0$ everywhere is a correct and necessary condition to enable a sensible calculation of the capacitance between two actual electrodes to be made, while the condition $\nabla \,.\, D_b \equiv 0$ alone prevents one from obtaining results of any significance.

We now have three ways of dealing with this problem:

(*a*) That illustrated by van der Ziel (Reference 2 in the paper), in which only the currents in the electrodes necessary to build up the surface charge on an arbitrarily chosen electrode are considered.

(*b*) That due to Llewellyn, in which only the currents necessary to satisfy the space-charge conditions are considered.

(*c*) That given in my paper, in which both currents are made to give a divergenceless total current between the electrodes.

Despite Rothe's measurements, I would like to point out that, since the theory given in my paper was worked out, Mr. Cotterhill of the College of Technology, Birmingham, showed me the results of measurements he had made a considerable time ago on the capacitance of a small diode. The capacitance varies with the operating conditions by very nearly the correct amount and certainly in the almost discontinuous way predicted in this paper. There are, however, very interesting differences which can be completely explained by the existence of the emission velocities of the electrons in an actual valve. It is to be hoped that Mr. Cotterhill will soon be able to publish his results.

* BULL, C. S., Paper No. 2333 R, July, 1957 (see 104 B, p. 374).

† e.g. ROTHE, H., *Proceedings of the Institute of Radio Engineers*, 1940, **28**, p. 325, Fig. 6.

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The Institution of Electrical Engineers
Paper No. 2416 R
Nov. 1957

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CHARACTERISTICS OF SOME FERROUS AND NON-FERROUS WAVEGUIDES AT 27 Gc/s

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(*The paper was first received* 1*st March, in revised form* 16*th May, and in final form* 5*th July*, 1957.)

SUMMARY

Measurements of the attenuations produced by air-filled, rectangular waveguides of brass, copper, aluminium, nickel-silver, nickel, mild-steel, Radiometal, Rhometal and Mumetal have been made at a frequency of 27 Gc/s, and the effective permeabilities of the ferrous materials at this frequency have been estimated from the results.

LIST OF PRINCIPAL SYMBOLS

a = Short internal dimension of waveguide.
b = Long internal dimension of waveguide.
E = Electric field strength.
H = Magnetic field strength.
K_{T1}, K_{T2}, K_p = Surface-roughness factors in attenuation formula.
γ = Propagation coefficient.
Z = Waveguide-wall wave impedance.
α = Attenuation produced by wall metal.
ϵ = Permittivity of dielectric inside waveguide (usually air).
ρ = Resistivity of waveguide wall metal.
σ = Conductivity of waveguide wall metal.
λ_e = Free-space wavelength.
λ_{cr} = Critical guide wavelength.
λ_g = Guide wavelength.
θ = Magnetic loss angle.
$\mu = \mu' - j\mu''$ = Complex permeability.
μ_R = 'Resistive' permeability.
μ_L = 'Inductive' permeability.

(1) INTRODUCTION

Losses in waveguides and cavity resonators, produced by the flow of induced currents in the metal walls, can be calculated from formulae derived on the assumption that the surfaces are perfectly smooth. It is now well known, however, that at microwave frequencies the measured losses are considerably greater than those calculated from such formulae, and it has been concluded,[1–3] from a series of investigations at frequencies about 10 Gc/s on drawn copper, brass and electroplated waveguides, that such discrepancies are due solely to surface roughness. The attenuations produced by waveguides of nickel, mild-steel, Mumetal, Radiometal and Rhometal have also been determined in the frequency range 9–9·675 Gc/s, and it has been shown that such measurements enable the effective high-frequency permeabilities of the materials to be estimated.[4]

The paper describes measurements which have been made on waveguides of several materials, both ferrous and non-ferrous, at 27 Gc/s.

(2) EXPERIMENTAL PROCEDURE

The attenuation measurements have been carried out by observing the voltage-standing-wave ratio in a short-circuited length of waveguide made from the material being studied. The commercially-pure-nickel sample and the four non-ferrous specimens were all drawn tubes. The mild-steel waveguide was carefully machined from the solid and joined down the centre of the long sides. The three nickel-iron-alloy guides were each fabricated from 0·015 in sheet with junctions along the centres of the long sides. The internal dimensions of all the waveguides were 1·068 cm × 0·431 cm.

The internal surface finish of each waveguide (except the aluminium one) was observed by the microscopic technique previously described,[3] to determine the magnitudes of the irregularities and to find the values of the coefficients* K_{T1}, K_{T2} and K_p. Similar coefficients were not obtained for aluminium because of the difficulty of plating it without changing the original surface.

(3) RESULTS

The attenuations produced by the various specimens at 27 Gc/s, together with the measured d.c. resistivities and the surface-roughness factors are given in Tables 1 and 2. The ratios between the measured and calculated attenuations are given for the non-ferrous waveguides, and details of the analysis and initial permeability of each ferromagnetic sample are also included.

(4) DISCUSSION OF RESULTS

The measured losses for the non-ferrous waveguides are all well in excess of the theoretical results as predicted by the Kuhn formula for perfectly smooth walls.[5] Measured values of the surface-roughness factors of the guide walls have been sufficient to account for losses in excess of theoretical figures, determined at frequencies near 10 Gc/s. The roughness factors recorded for the waveguides under consideration are not nearly sufficient, however, to account for the large increases. The principal explanation of this is thought to lie in the difference between the type of roughness measured and the nature of the roughness existing in practice. A cross-section of a conducting surface will, in general, look similar to that shown in Fig. 1 and will not be a perfect plane but will have irregularities. Now, under low-frequency conditions the current flow occupies a deep layer of the material, as shown in Fig. 1(*a*). The irregular surface line will therefore have little or no effect on the electrical properties of the conductor, since the irregularities are small compared with the skin depth. When the frequency is increased and the skin depth becomes small, however, the effective length of the surface is increased, owing to the tendency of the current flow to follow the boundary line [Fig. 1(*b*)]. This kind of fairly-smooth undulating surface can be measured without much difficulty and the corresponding roughness factors can be obtained. This is, in fact, what has been done to obtain the figures in Table 1. The surfaces examined here had irregularities of very small dimensions—indeed, such that it was impossible to measure the true lengths of the surfaces accurately [Fig. 1(*c*)]. The fine roughness under discussion is thought to be caused by the drawing dies, which are presumably difficult to polish in small sizes. It

Written contributions on papers published without being read at meetings are invited for consideration with a view to publication.
Dr. Allison and Dr. Benson are at the University of Sheffield.
Mr. Seaman was formerly at the University of Sheffield and is now with Pye, Ltd.

* For definitions of K_{T1}, K_{T2} and K_p, see Reference 3.

Table 1

CHARACTERISTICS OF NON-FERROUS WAVEGUIDES

Material	D.C. resistivity	Measured attenuation at 27 Gc/s	K_{T1}	K_{T2}	K_p	Measured attenuation / Calculated attenuation	
						Neglecting roughness†	Including roughness factors*
	microhm-cm	dB/m					
Brass	6·40	0·78	1·047	1·046	1·024	1·24	1·20
Copper	1·77	0·46	1·055	1·056	1·027	1·39	1·34
Aluminium	2·98	0·60	—	—	—	1·39	—
Nickel-silver	30·2	1·61	1·002	1·003	1·006	1·18	1·16

* Calculated values obtained from the expression:[3]

$$\alpha = 0{\cdot}159\,(\rho\mu_R)^{1/2}\frac{\lambda_g}{b(\lambda_e)^{3/2}}\left[\left(K_{T2}+\frac{b}{2a}K_{T1}\right)\frac{\lambda_e^2}{\lambda_{cr}^2}+K_p\left(1-\frac{\lambda_e^2}{\lambda_{cr}^2}\right)\right] \text{ decibels per metre}$$

where the wavelengths and guide dimensions are in centimetres and ρ is in microhm-centimetres.

† Calculated values obtained from the above expression with $K_{T1} = K_{T2} = K_p = 1$, i.e. the Kuhn formula.[5]

Table 2

CHARACTERISTICS OF FERROMAGNETIC WAVEGUIDES

Material	Chemical analysis	D.C. resistivity	Initial permeability	Measured attenuation at 27 Gc/s	K_{T1}	K_{T2}	K_p	μ_R as calculated from equation at foot of Table 1
		microhm-cm		dB/m				
Nickel	Contains following impurities: Fe = 0·58% S = 0·022% Mn = 0·12% Si = Trace	8·59	13·5	0·77	1·026	1·038	1·013	1·1
Mild steel ..	Contains following impurities: C = 0·27% S = 0·053% Mn = 0·74% Si = 0·03% P = 0·036% Ni = 0·15%	13·5	141	2·46	1·049	1·078	1·018	6·7
Mumetal ..	Ni = 72·2%* Cu = 5·3%	65·3	21 000†	1·93	1·002	1·002	1·001	0·9
Radiometal ..	Ni = 49·1%* Cu = 0·0%	46·6	1 850†	2·47	1·002	1·002	1·001	2·1
Rhometal ..	Ni = 37·5%* Cu = 0·0%	74·2	1 200†	2·91	1·002	1·002	1·001	1·8

* Remainder iron with traces of molybdenum and chromium.
† Manufacturer's figures.

is interesting to note that irregularities of this order have not normally been found in waveguides for use at 3 cm.

The theoretical treatment by Morgan[6] predicts an increase of approximately 50–60% for the case of a surface in which the roughness is comparable with the skin depth. Although the special kind of surface studied by Morgan is not found in practice, and his results would be expected to be higher than experimental values obtained for surfaces with random irregularities, it does seem that minute roughness of the surface is the principal source of the discrepancy which has just been discussed. As pointed out by Morgan, however, roughness parallel to the current flow in the walls will produce an increase in attenuation which cannot be calculated other than by approximate methods, and in the derivation of the equation at the foot of Table 1 such effects have been entirely neglected. It is also possible that the resistivity of the actual surface layer, which must be highly strained, is greater than that of the general bulk of the waveguide metal on which d.c. measurements were made to determine the resistivity. In addition, tarnish films will account for some loss.

Nickel-silver waveguide has much greater attenuation than drawn copper, brass or aluminium, owing to its high resistivity. Since it also has corrosive properties similar to those of the other waveguides, its only advantage seems to be its greater mechanical strength.

It is seen from Table 1 that the measured attenuations for non-ferrous materials are 16–34% greater than the corresponding calculated values including roughness factors. Such discrepancies have been ignored in calculating the permeability of each ferromagnetic waveguide. It might therefore be argued that the

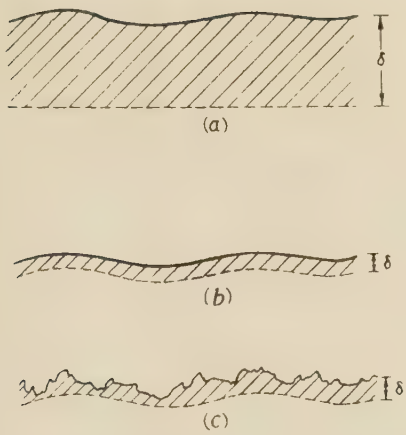


Fig. 1.—Sections of conducting surfaces showing the various types of roughness.

(*a*) Irregularities small compared with skin depth, δ.
(*b*) Skin depth small; current follows boundary line.
(*c*) Irregularities of very small dimensions.

figures given for μ_R in Table 2 are from 30 to 80% too large. The ferromagnetic waveguides, only one of which was a drawn tube, did not have the same microscopic roughness on their surfaces, however, as the non-ferrous waveguides, and in the case of the nickel-iron alloys which have high resistivities, the skin-depths will be relatively large. The errors in the values of μ_R calculated are thus thought to be much less than the percentages quoted above.

Several alternative theories to explain the phenomenon of the dispersion of apparent permeability with increase of frequency have been briefly discussed previously,[4] but many do not account satisfactorily for the dispersion in ferromagnetic materials above 200 Mc/s, although they may be appropriate at lower frequencies.[7] The observed dispersion at high frequencies is closely associated with the depth of penetration of the field being comparable with the thickness of the ferromagnetic domains, estimated to be between 10^{-3} and 10^{-4} cm. It also seems likely that the damping term associated with spin precession in ferromagnetic-resonance experiments is an important factor.[8,9] All these theories indicate that the magnetic permeability should approach unity at very high frequencies. It seems that the frequency at which the permeability becomes unity is almost reached at 27 Gc/s for all the materials examined except mild steel. Certainly, at wavelengths of the order of 0·5 cm, these four metals should behave as though they were non-magnetic. The apparent permeabilities

Table 3

PERMEABILITIES OF THE VARIOUS MATERIALS

Material	μ_R	
	At 9·675 Gc/s	At 27 Gc/s
Nickel	3·5	1·1
Mild steel	45	6·7
Mumetal	3·0	0·9
Radiometal	12·0	2·1
Rhometal	11·0	1·8

of the materials at frequencies of 9·675 and 27 Gc/s are listed in Table 3.

The permeability of a specimen at high frequencies may be determined by methods which fall into two separate groups.[4] In the first, resistive losses in an arrangement containing the ferromagnetic material are measured; the permeability deduced from the results is called the resistive permeability, μ_R, and it is this quantity which has been measured during the present investigations. In the second, the permeability is obtained from the effective reactance of a similar arrangement, this being designated μ_L. It is shown in the Appendix that the following simple relationships can be found connecting μ_R, μ_L and the complex permeability $\boldsymbol{\mu} = \mu' - j\mu''$:

$$\mu_R = \sqrt{[(\mu')^2 + (\mu'')^2]} + \mu'' \quad . \quad . \quad . \quad (1)$$

and

$$\mu_L = \sqrt{[(\mu')^2 + (\mu'')^2]} - \mu'' \quad . \quad . \quad . \quad (2)$$

from which

$$\mu'' = \tfrac{1}{2}(\mu_R - \mu_L) \quad . \quad . \quad . \quad . \quad . \quad . \quad . \quad (3)$$

and

$$\mu' = \sqrt{(\mu_R \mu_L)} \quad . \quad . \quad . \quad . \quad . \quad . \quad . \quad . \quad (4)$$

It follows that, in order to determine the complex permeability, $\boldsymbol{\mu}$, of a material, μ_R and μ_L must be measured at the same frequency. Dispersion in μ_L occurs, however, at frequencies well below those at which μ_R variation is greatest. In fact, investigations by Millership and Webster[10] showed that, for iron, steel and Mumetal, μ_L was unity at 1·5 Gc/s.

(5) ACKNOWLEDGMENTS

The work recorded in the paper has been carried out in the Department of Electrical Engineering at the University of Sheffield. The authors wish to thank Professor A. L. Cullen for facilities afforded in the laboratories of this Department and for his encouragement, advice and helpful suggestions. They are also grateful to Mr. H. B. Johnson, for carrying out some of the attenuation measurements on ferromagnetic waveguides.

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(7) APPENDIX

Relationships between μ_R, μ_L and the components of μ

Maxwell's equations give

$$\text{curl } \boldsymbol{E} = -j\omega\mu\boldsymbol{H} \quad . \quad . \quad . \quad . \quad . \quad (5)$$

and

$$\text{curl } \boldsymbol{H} = (j\omega\epsilon + \sigma)\boldsymbol{E} \quad . \quad . \quad . \quad . \quad . \quad (6)$$

Suppose that only components E_x and H_y exist and write

$$\mu = \mu' - j\mu'' = |\mu|\varepsilon^{j\theta}$$

Eqns. (5) and (6) then give

$$\frac{\partial E_x}{\partial z} = -j\omega(\mu' - j\mu'')Hy \quad . \quad . \quad . \quad . \quad . \quad (7)$$

and

$$\frac{\partial H_y}{\partial z} = (\sigma + j\omega\epsilon)E_x \quad . \quad . \quad . \quad . \quad . \quad . \quad (8)$$

If the dependence of the fields on the z-co-ordinate is of the form $\varepsilon^{-\gamma z}$, eqns. (7) and (8) give

$$-\gamma E_x = (-j\omega\mu' - \omega\mu'')H_y \quad . \quad . \quad . \quad . \quad . \quad (9)$$

and

$$\gamma H_y = (\sigma + j\omega\epsilon)E_x \quad . \quad . \quad . \quad . \quad . \quad . \quad (10)$$

Dividing eqn. (9) by eqn. (10) gives

$$\frac{E_x}{H_y} = \left[\frac{(j\mu' + \mu'')\omega}{\sigma + j\omega\epsilon}\right]^{1/2} \quad . \quad . \quad . \quad . \quad . \quad (11)$$

Now

$$\sigma \gg \omega\epsilon$$

Therefore

$$\frac{E_x}{H_y} \simeq \left[\frac{\omega(\mu'' + j\mu')}{\sigma}\right]^{1/2} \quad . \quad . \quad . \quad . \quad . \quad (12)$$

The loss in the waveguide depends on the real part of this expression.

Now

$$\begin{aligned}\omega(\mu'' + j\mu')^{1/2} &= (j\omega|\mu|\varepsilon^{-j\theta})^{1/2} \\ &= (\omega|\mu|)^{1/2}(j^{1/2}\varepsilon^{-j\theta/2}) \\ &= (\omega|\mu|)^{1/2}\exp\left[j\left(\frac{\pi}{4} - \frac{\theta}{2}\right)\right] \\ &= (\omega|\mu|)^{1/2}\left[\cos\left(\frac{\pi}{4} - \frac{\theta}{2}\right) + j\sin\left(\frac{\pi}{4} - \frac{\theta}{2}\right)\right] \quad (13)\end{aligned}$$

Thus, for the waveguide-wall wave impedance,

$$Z = \frac{E_x}{H_y} = \left(\frac{\omega|\mu|}{\sigma}\right)^{1/2}\left[\cos\left(\frac{\pi}{4} - \frac{\theta}{2}\right) + j\sin\left(\frac{\pi}{4} - \frac{\theta}{2}\right)\right] \quad (14)$$

For a hysteresis-free material, $\theta = 0$ and

$$Z = \left(\frac{\omega|\mu|}{\sigma}\right)^{1/2}\frac{(1+j)}{\sqrt{2}} \quad . \quad . \quad . \quad . \quad (15)$$

If μ_R is defined as that value of $|\mu|$ which when substituted in eqn. (15) would give the same real part of Z as the specimen, then μ' and μ'' are related to μ_R as follows:

$$|\mu|\cos^2\left(\frac{\pi}{4} - \frac{\theta}{2}\right) = \frac{\mu_R}{2} \quad . \quad . \quad . \quad . \quad (16)$$

i.e.

$$\frac{\mu_R}{2} = \frac{|\mu|(1 + \sin\theta)}{2} \quad . \quad . \quad . \quad . \quad . \quad (17)$$

But

$$|\mu|\sin\theta = \mu'' \text{ and } |\mu| = [(\mu')^2 + (\mu'')^2]^{1/2}$$

so

$$\mu_R = \sqrt{[(\mu')^2 + (\mu'')^2]} + \mu'' \quad . \quad . \quad . \quad (18)$$

Similarly, if μ_L is defined as that value of $|\mu|$ which when substituted in eqn. (15) would give the same imaginary part of Z as the specimen,

$$\frac{\mu_L}{2} = |\mu|\sin^2\left(\frac{\pi}{4} - \frac{\theta}{2}\right) \quad . \quad . \quad . \quad . \quad (19)$$

which gives

$$\mu_L = \sqrt{[(\mu')^2 + (\mu'')^2]} - \mu'' \quad . \quad . \quad . \quad (20)$$

Eqn. (18) can also be easily obtained by starting with the expression given by Kuhn,[5] in calculating attenuation, for the amount of power ΔP passing into the waveguide walls in a length Δx, then introducing the complex permeability and determining the real part of $\Delta P/\Delta x$.

DISCUSSION ON

'THE CONTROL AND INSTRUMENTATION OF A NUCLEAR REACTOR'

NORTH-EASTERN CENTRE, AT NEWCASTLE UPON TYNE, 12TH NOVEMBER, 1956

Mr. A. G. Entwistle: When a reactor is operated well into the sub-critical region, it is only feasible to control it by reference to instruments presenting information relating to neutron flux, since the heat outputs are too small to enable other types of instrument to detect the onset of a dangerous power rise. However, at power-generation levels, neutron-flux measurements seem to be used fundamentally to determine primary heat output by a method which differs from that normally adopted in power stations. Does the author consider it equally practicable at power levels to control a reactor by reference to normal industrial instruments measuring the heat output?

A recent paper by Cox and Sandiford describes the wire-activation method of measuring the radial and longitudinal distribution of neutron flux employed at Calder Hall. This method can readily be applied to reactors of large physical dimensions, but it could not be so easily adapted to the small and highly rated reactors at present in the research stage. What methods are used for the measurement of flux distribution in physically small reactors?

The methods of control described in the paper are applicable to designs of reactor in which the effective reactivity depends principally on control-rod setting. However, in the boiling-water reactor an increase in heat generation will result in an increase in the volume of steam in the reactor core and a corresponding decrease in the volume of water. Since the water acts as

* Gillespie, A. B.: Paper No. 2058 M, March, 1956 (see 103 B, p. 564).

moderator as well as coolant, the proportion of neutrons being reduced to thermal energy decreases, so tending to reduce the reactor power. These reactors appear to be largely self-regulating, and raise a control problem different from those discussed in the paper: will the author enlarge upon the design of control and safety systems envisaged for them?

Mr. T. H. Morrison: To a power engineer, a few years ago it would have seemed fantastic that currents of the order of 10^{-12}amp, mentioned in the paper, could be measured, let alone be used for practical purposes. However, we are becoming more familiar with them and the associated cables, although there is some doubt about the amount of interference picked up from neighbouring power and communication circuits. It was once general practice to ensure that there was adequate clearance, 6ft or more, around special control and instrumentation cables, but it appears that this is now being relaxed. Present practice seems to bundle these cables into metal trunking which can be run in close proximity to all circuits. It would be of interest to have the author's views on this point and also on the make-up of special cables which have been developed for these circuits.

It is not clearly indicated in the paper where the ionization chambers are positioned in a reactor. In a power reactor the chambers are located in the thermal column, which is external to the pressure vessel. How are the instruments calibrated for such locations, and what is their accuracy compared to those within the pile?

Mr. E. J. M. Marrian: In the control of this experimental reactor emphasis seems to have been placed on shutting it down if anything begins to go wrong, rather than on keeping it going as long as possible. If excess neutron flux is indicated, is it really necessary to shut it down immediately, or is there time, before run-away, to bring it under control? This would be of paramount importance in a commercial reactor which was relied upon to maintain national power supplies.

I think much could be done to improve the protective circuits shown in Fig. 11. For instance, a battery supply free of earth (or earthed through a high resistance to give earth-fault indication) would lessen the risk of shut-down due to stray earthing. Furthermore, there does not seem to be much point in including the auxiliary relays C, D, E, F, etc., in the shut-down circuit. It would be better, I think, to use duplicate contacts of the excess-flux and low-period indicators to give separate operation of relays A and B, leaving the auxiliary relays to give audible and visual alarms.

Mr. T. Sealy: It has become standard practice to earth the positive pole of 50-volt battery supplies, but this convention has not been followed in the circuit shown in Fig. 11. The type of relay used is not indicated, but I hope that the standard Post Office 3000-type relay has been employed, for it has proved superior to many so-called 'specially designed' relays in this type of circuit.

The author states that relays A and B monitor the safety line at the top and the bottom. This is unnecessary, since it is a series circuit, and with the relay coils connected in one side only of the circuit, any open-circuit will de-energize the relays. The arrangement shown has practical disadvantages, namely that the coil of relay A is always at a potential of 50 volts with respect to earth, with the possibility that troubles may arise from electrolysis.

The circuit operating relay G must be regarded with suspicion, since, in practice, 20 or more of these excess-flux contacts can be connected in series. With the circuit shown, one faulty contact will shut down the reactor. Any relays used must have sufficient contacts to allow them to be connected in paralleled pairs and ensure a much more reliable circuit.

Alarm lamps are shown supplied from a separate a.c. supply, presumably to prevent excessive battery drain. This circuit will, I assume, be arranged with a.c./d.c. change-over to enable the lamps to be supplied from the battery in an emergency.

The relays are described as being of the non-self-resetting type. In a practical control room it would be more convenient to have self-resetting relays, with separate relays provided for the alarm circuits, locking-in electrically so that the alarm could be given to control, accepted by the control engineer, logged and cleared from a single pushbutton on his desk, instead of resetting the local relay.

Mr. C. H. Morton: In Section 3 it is stated that an artificial neutron source is inserted to reduce the range of neutron-flux measurement required, and I wonder whether the radioactivity of the fission products during service increases the effectiveness of this source and so decreases still further the instrumentation range. In other words, is the range required constant throughout the useful life of the reactor?

In Fig. 2, excess fuel temperature is shown as giving a warning, while excess coolant temperature shuts down the reactor. I should have thought that the fuel temperature, which must give an earlier indication of trouble, would be the more suitable measurement for an emergency shut-down.

Finally, what is the practical significance of reactor low period, and why is it necessary for it to shut down the reactor as indicated in Fig. 2?

Mr. G. H. Hickling: The instrumentation described in the paper is of great complexity, many types of neutron detectors and of associated amplifiers being employed, for example, to observe and control the neutron flux from start-up to power-operating conditions. Although this may be necessary and quite acceptable on the research reactors with which the author is concerned, it is desirable that for future nuclear power reactors a large measure of simplification should be effected. With this in view, what does the author regard as the minimum requirements for direct nuclear measurements in the new piles?

Little mention is made in the paper of the effect of temperature coefficients on reactor stability. I believe that this contributes materially to reactor safety and ease of control. Can the author state the contribution to the temperature coefficient of reactivity made by the graphite, and whether these factors can be made use of directly in controlling a reactor at power? With reference to Fig. 9, I should be interested to know the maximum power rating to which the method shown for induction-motor speed control has so far been applied.

In all the reactors referred to in the paper, safety has been made very greatly dependent on electronic circuits. Apart from the practice of duplication of essential circuits, would the author say what measures are currently adopted to ensure the necessary high degree of reliability in the electronic devices themselves? (For example, 'soak testing' has, I believe, been used to cut down premature failures.) Are germanium diodes and transistors or magnetic amplifiers employed to any extent in place of thermionic valves?

Can the author clarify the statement, in discussing 'burst slug' detection, that gaseous isotopes of xenon and krypton can be isolated by filtering the coolant, and also explain the scanning process by which individual defective elements can be located? Would he also give some indication of the circumstances, or the locations in the pile, in which it becomes necessary to use pulse counters to differentiate between neutrons and γ-rays, and since the BF_3 type of pulse counter is shown to be more sensitive than the fission counter, would he explain the stated preference for the latter in power reactors?

Mr. A. B. Gillespie (*in reply*): In reply to Mr. Entwistle, thermal instruments measuring the heat output of the reactor in terms of coolant flow and temperature rise are essential when the reactor

is in the power range, and such information would be used as the basis for the longer-term control of the reactor. For the short-term control it is necessary to refer to the neutron-flux measuring instruments, because these act quickly and the neutron flux is a direct measure of the state of the fission chain reaction. The thermal output is not directly proportional to the fission reaction rate, because of the slowly varying contribution from the build-up and radioactive decay of fission products.

Other methods of flux scanning use small fission counters in the range 10^4–10^8 n/cm^2/sec and boron-coated thermocouples from 10^{10} n/cm^2/sec upwards.

The basic methods of control and safety outlined in the paper should be applicable to the boiling-water reactor. Although such a reactor is inherently self-regulating at a given power level, the same facilities for power alteration and safety are required as in the conventional graphite reactors, with alterations only in detail to suit the different mechanics and dynamic characteristics of the reactor.

In reply to Mr. Morrison, the signal leads carrying currents of 10^{-12} amp are of the polythene-insulated Uniradio series. Low-microphony cables are necessary, and are available commercially with a thin semi-conducting plastic film between the polythene and the metal braid to prevent build up of charge on the polythene surface due to cable flexing or vibration. Signal leads are prone to interference, and care must be taken to avoid this. In general, all low-level signal leads should be separated from power-supply cables and control cables to motors and relays; they should have their own trunking and be spaced as far away as is practicable. Continuity of earth through the trunking, as well as good earthing of the trunking cover, is very important.

The siting of ionization chambers is dealt with in detail in the companion paper by Abson and Wade.*

I agree with Mr. Marrian that in a power reactor the emphasis should be on not shutting it down until absolutely necessary. This involves duplication of safety circuits and instruments to guard against accidental shut-down due to instrument failure, together with the grouping of fault conditions into categories in which less drastic action may be taken in all but the most serious. Such a system is incorporated in most reactors now under construction. Fig. 11 is not intended to represent the last word in safety-circuit design; it was included to give the uninitiated reader a simple picture of one form which such circuits may take. The design of these circuits is the subject of much attention, and considerable progress towards increased safety and reliability has been achieved since the preparation of Fig. 11.

In reply to Mr. Morton, the use of artificial neutron sources is discussed in the companion paper by Cox and Walker.†

In Fig. 2 excess fuel temperature should be in the shut-down column and excess coolant temperature in the warning column.

In the power range from shut-down to a level where temperature coefficients are having an effect, the reactor power rises exponentially. The period is the time taken for the power to increase by a factor ε and is constant for a given exponential rise. It is therefore convenient to measure this parameter, since this effectively defines the rate at which the power is increasing, and to use this information to bring about an early shut-down if the power is rising too fast. This is important because the normal level shut-down devices do not usually operate until an appreciable power level has been reached, and if the power is rising rapidly the shut-down they cause may be accompanied by a damaging power overshoot.

I do not agree with Mr. Hickling that the neutron-flux instruments listed in the paper are complex. In a small zero-energy reactor, where they may form the bulk of the instrumentation, this may appear so, but the justification is the experimental nature of the reactor and its use. On a power reactor the nuclear instruments form a much smaller part of the total instrumentation, and the neutron-flux measuring instruments might be reduced to a logarithmic d.c. amplifier with period indication for the start-up range and a linear d.c. amplifier with backing-off for error indication for the power range. It is likely, however, that one set of pulse-counting equipment would also be necessary during shut-down, and duplicate measuring instruments are advisable in the interests of safety and reliability.

The temperature coefficients of both the fuel and the moderator contribute materially to the stability of a graphite-moderated reactor in the power range. The advantages of these two factors, which together amount to about $-2{\cdot}8 \times 10^{-5}$ in reactivity per degC, are fully exploited by reactor designers, and Fig. 5 in the paper by Cox and Walker* shows markedly how they limit the rate of power increase.

The motor used in the speed-control system described in the paper had a rated output of 0·1 h.p., but if magnetic amplifiers are used to supply power to the stator windings, the system should be suitable for much higher powers.

To secure instrument reliability, all the usual precautions are taken such as using generously rated components and providing efficient and adequate ventilation. Valve failures are the most troublesome, and since most of those occur in early life, it is standard practice to soak-test all instruments for some 200 hours prior to use. Adequate spares are also carried, so that a faulty instrument can be replaced immediately. Transistors have not yet come into general use in reactor instruments, but a preliminary exploration of the field indicates that most instrument designs can be converted fairly easily and that improved reliability should result. Magnetic amplifiers have been used where power requirements are outside the capabilities of thermionic valves, but have not found much use in the low-power measuring and safety instruments because of their relatively long time-constants.

The neutron/γ ratio in a reactor is usually most unfavourable a short time after shutting down a reactor which has been running at power, owing to the high γ background set up by the fission products in the reactor fuel, and pulse-counting detectors are usually necessary to secure adequate neutron to γ discrimination. Fission counters are generally preferred in power reactors because relatively high neutron fluxes are available to offset their reduced neutron sensitivity; they also have a longer life and better neutron/γ discrimination than BF_3 counters, which are more frequently used when a reactor is being loaded with fuel for the first time, for fluxes are then low and maximum neutron sensitivity is required.

* Abson, W., and Wade, F.: 'Nuclear-Reactor-Control Ionization Chamber', *Proceedings I.E.E.*, Paper No. 2029 M, March, 1956 (103 B, p. 590).

† Cox, R. J., and Walker, J.: 'The Control of Nuclear Reactors', *ibid.*, Paper No. 2068 M, March, 1956 (103 B, p. 577).

* Cox, R. J., and Walker, J.: 'The Control of Nuclear Reactors', *ibid.*, Paper No. 2068 M, March, 1956 (103 B, p. 577).

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ABBREVIATIONS

(P)—Address, lecture or paper.
(p)—Subject dealt with in a paper or address.
(D)—Discussion on a paper.
(A)—Abstract of paper or address.

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